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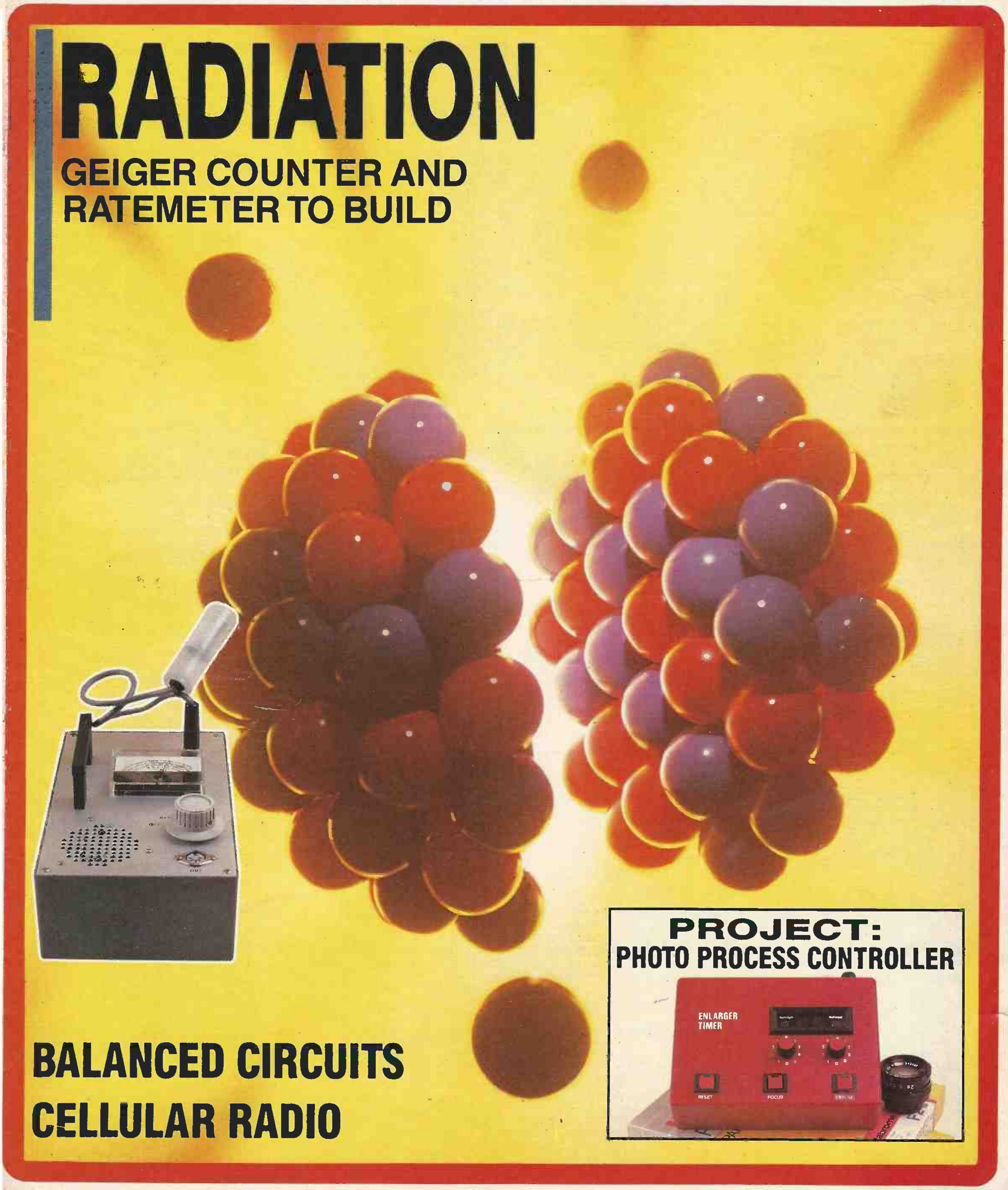

RADIATION

GEIGER COUNTER AND RATEMETER TO BUILD

PROJECT:
PHOTO PROCESS CONTROLLER

BALANCED CIRCUITS

CELLULAR RADIO

AUDIO....COMPUTING....MUSIC....RADIO....ROBOTICS..

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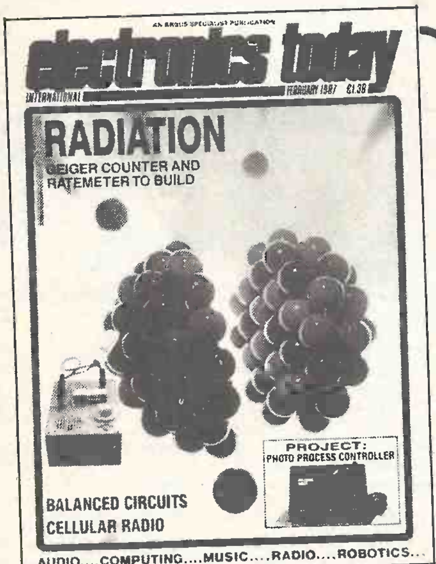

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FEATURES

A QUESTION OF BALANCE 14
Barry Porter tips the scales in favour of noise-free signals with the low down on balanced lines for stage and home audio systems.

HARDWARE DESIGN CONCEPTS 21
Mike Barwise takes his design ideas to the workbench and starts to build a top specification pulse generator. We're all invited to watch.

CELLULAR RADIO 25
The top telephones of the rich come under scrutiny. Keith Brindley 'phones in his report from a traffic jam on the M1.

THE MIDI INTERFACE 31
This month Alan Robinson takes a look at the control codes used by the Musical Instrument Digital Interface standard.

PROJECTS

GEIGER RATEMETER AND COUNTER 35
A general purpose radiation detector to give an accurate reading for anything from a lump of uranium to some dodgy vegetables. Colin Seymour presents this precision instrument.

PHOTO PROCESS CONTROLLER 41
Ian Coughlan's photographic enlarger timer is the best design around. With the quality of an SLR and the simplicity of an Instamatic, it's a snap.

LOW DISTORTION STEREO DECODER 46
John Linsley Hood turns from the theory of the past few months to a practical high quality FM stereo decoder to add to an existing tuner or to keep for his tuner project coming next month.

THE ETI LEDLINE ANALYSER... 50
A fully fledged logic analyser is beyond the reach of most overdrafts. The LEDline analyser does the same job for a fraction of the price.

REGULARS

NEWS DIGEST 6
All that's happened.

DIARY 9
All that's going to happen.

READ/WRITE 13
More of your letters. What's the point of MIDI? What's the future of records? What's the price of a second hand oscilloscope? It's all here.

TECH TIPS 54
More of your circuits, and with a decidedly automotive flavour. For those with tow-bars there's a trailer light controller and a 50Hz inverter for the caravan.

INFORMATION

NEXT MONTH'S ETI 8
SUBSCRIPTIONS 11
SPECIAL OFFER 30
PCB SERVICE 56
PCB FOILS 59
FREE READERS' ADS 62
CLASSIFIED ADS 64

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LF355	80p
LF356N	110p
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TA7310	150p
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TBA820M	75p
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TDA2030	250p
TDA2593	300p
TDA3810	750p
TDA7000	300p
TEA1002	700p
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TL062	90p
TL064	90p
TL071	40p
TL072	75p
TL074	110p
TL081	30p
TL082	60p
TL083	70p
TL084	100p
TL094	180p
UA759	300p
UA[illegible]	150p
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VOLTAGE REGULATORS

FIXED PLASTIC

1A		+ve		+ve	
5V		7805	45p	7905	50p
6V		7806	50p	7906	50p
8V		7808	50p	7908	50p
12V		7812	45p	7912	50p
15V		7815	50p	7915	50p
18V		7818	50p	7918	50p
24V		7824	50p	7924	50p
5V	100mA	78L05	30p	79L05	45p
8V	100mA	78L08	30p		
12V	100mA	78L12	30p	79L12	50p
15V	100mA	78L15	30p	79L15	50p

OTHER REGULATORS

Fixed Regulators		
LM309K	1A 5V	140p
LM323K	3A 5V	250p
78H05KC	5A 5V	575p
78H12	5A 12V	750p
Variable Regulators		
LM305AH		250p
LM317T	TO-220	120p
LM317K	TO3	240p
LM337T	3A+VAR	225p
LM350T	5A+VAR	400p
LM396K	10A+VAR	£15
LM723N		50p
78HGKC	5A+VAR	650
79HGKC	5A+VAR	675p
78GUIC	1A+VAR	225p
79GUIC	1A+VAR	250p
Switching Regulators		
ICL7660		250p
SG3524		300p
TL494		300p
TL497		225p
78S40		250p

COMPUTER COMPONENTS

CPU

1802CE	650p
2650A	1050p
6502	400p
6502A	450p
65C02A	£15
6800	250p
6802	300p
6809	650p
6809E	£8
68B09	£10
68B09E	£12
68000-LB	£36
8035	350p
8039	420p
80C39	700p
8080A	750p
8085A	300p
8087	120p
80C85A	900p
8086	£22
8088	1750p
8741	£12
8748	£12
TMS9980	£14.50
TMS9995	£18
Z80	250p
Z80A	280p
Z80B	550p
Z80CMOS (CMOS Z80)	750p
808A	850
8279	480p
8282	300p
8284	400p
8287	300p
8288D	650p
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Z80DART	600p
Z80ADART	700p
Z80DMA	700p
Z80ADMA	750p
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Z80BCTC	600p
Z80DART	600p

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2816-30	£20

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6829	£12.50
6840	375p
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6850	180p
68B50	300p
6852	300p
6854	600p
68B54	600p
6875	300p
8154	600p
8155	300p
8156	300p
8205	225p
8212	300p
8216	180p
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8226	425p
8228	600p
8243	300p
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2532-30	600p
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2732	400p
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ULN2803	180p
ULN2804	180p
75107	90p
75108	90p
75109	120p
75110	90p
75112	180p
75113	120p
75114	140p
75115	140p
75121	140p
75122	140p
75150P	130p
75154	130p
75159	220p
75160	[illegible]
75161	650p
75162	750p
75172	400p
75182	150p
75188	60p
75189	60p
75365	180p
75450	90p
75451	60p
75452	70p
75453	70p
75454	70p
75480	180p
75491	60p
75492	60p
8T26	120p
8T28	120p
8T95	120p
8T96	120p
8T97	120p
8T98	120p

81LS95	140p
81LS96	140p
81LS97	140p
81LS98	140p
88LS120	450p
9602	300p
9636A	140p
9637AP	180p
9638	180p
ZN425E8	350p
ZN426E8	350p
ZN427E	600p
ZN428E8	450p
ZN429E	210p
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4164-15	150p
4416-15	300p
4532-20	250p
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5514	450p
5516	400p
5517	400p
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1.00MHz	270p
Freq in MHz	
1.8432	225p
2.00	250p
2.45760	200p
2.5	200p
3.12MHz	175p
10.00MHz	175p
3.276	180p
3.5795	100p
4.00	140p
4.194	190p
4.43	100p
4.608	200p
4.9152	200p
5.000	150p
6.00	140p
17.734	200p
7.00	190p
7.168	175p
8.00	150p
8.867	175p
10.50	200p
10.70	150p
11.00	300p
12.00	190p
14.00	175p
14.318	180p
14.756	250p
15.00	200p
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116	250p
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20 pin	45p		

LOW PROFILE SOCKETS BY TEXAS

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14 pin	10p	24 pin	24p
16 pin	11p	28 pin	28p
18 pin	16p	40 pin	30p
20 pin	18p		

WIRE WRAP SOCKETS BY TEXAS

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18 pin	42p	28 pin	100p
18 pin	50p	40 pin	130p
20 pin	60p		

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TIL31A	120p	TIL81	120p
TIL100	120p	SFH305	100p

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74C926	650p
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		6N137	360p

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Green (10)	225p
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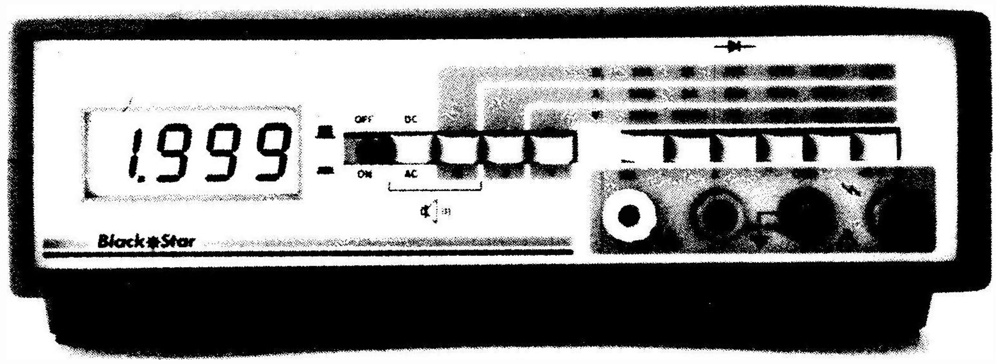
● The Botswana Technology Centre have written to tell us that their current electronics engineer is leaving in March 1987 and they are looking for a replacement. The work is aimed mostly at creating jobs and aiding rural development. Applicants with a degree in electronics and experience of design and production should contact the Recruitment Executive, ODA, Abercrombie House, Eagle Sham Road, East Kilbride, Glasgow G75 8EA.

● Several ranges of Oak switching products are described in two new catalogues from Diamond H Controls. Keyboards, keyboard switches, membrane panels and programmable rotary switches are covered in one catalogue while the other lists a wide range of rotary switches including PCB-mounting, lever-operated and miniature and subminiature types. Diamond H Controls Ltd, Vulcan Road North, Norwich NR6 6AH, tel (0603) 45291.

● A survey taken at the Internepcon exhibition in October reveals that 36% of those visiting the show are already using surface-mount technology and a further 15% are likely to adopt the technique in the coming year. The organisers of the exhibition claim that their visitors represent a broad cross-section of the UK electronics industry, and suggest on the basis of these findings that around 50% of all electronics manufacturers will be using surface mount techniques in 12 months time.

● The range of DIN 41612 PCB connectors manufactured by Bulgin is described in a new 16-page catalogue. The types featured include B, C, D, F, H and R connectors with from 11 to 96 contacts and with pin spacings from 0.1" to 0.4". Included in the catalogue is a selection guide, a series of dimension drawings plus board and panel hole-cutting patterns. A.F. Bulgin & Company PLC, Bypass Road, Barking Essex IG11 0AZ, tel (01) 594-5588.

● Electrovalue, whose new catalogue we mentioned in our December issue, will no longer be sending out each new issue automatically to past customers. Instead, those who require a copy will have to write and request one. However, so that no extra cost is involved for the customer, Electrovalue will be using a Freepost address which removes the need for a stamp on the envelope. Write to Electrovalue Ltd, Freepost, 28 St. Judes Road, Egham Surrey TW20 8BR or telephone (0784) 33603. The catalogue itself is still free.



Measuring Time

The latest digital multimeter from Black Star combines a wide range of measuring functions and high accuracy with a typical battery life which is equivalent to over ten months of continuous operation.

Designed as a bench instrument which is none the less extremely portable, the 3225 can operate for around 7,500 hours on one set of alkaline batteries. It has a 3½ digit display, a basic DC accuracy of ±0.25% and includes a diode test function and an audible continuity tester. AC and DC voltages are covered in five ranges with FSDs from 200mV to 1000V (750V AC) while resistance is covered in six ranges from 200R to 20M FSD. AC and DC currents are covered in six ranges from 200uA to 10A FSD and the AC measurements are average responding, RMS calibrated.

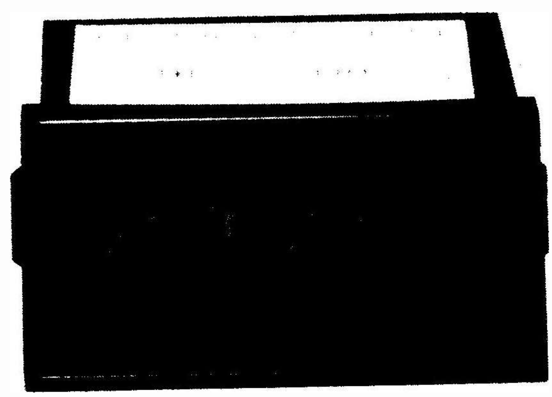
The liquid crystal display has 0.5" high characters and includes over-range, polarity and low-battery indication. Connection is made via recessed 4mm sockets and the moulded plastic case is shielded against RF and other electromagnetic interference. The 3225 comes complete with test leads, a set of batteries and a comprehensive manual and costs £89.00 plus VAT. The 3210, an upgraded version of the 3225 offering 0.1% DC accuracy and true-RMS AC measurements, will be available soon at £119.00 plus VAT.

Also new from Black Star is the Jupiter 500 function generator which offers sine, square and triangle waveforms over the frequency range 0.1Hz to 500kHZ. Two 600R outputs are provided, one of which delivers up to 30V and can be offset within the range −15V to +15V, and there is also a TTL-level square wave output capable of driving up to 30 standard loads. Coarse and fine frequency controls are provided and all outputs are protected against short circuits. External input signals can be used to control the output amplitude over its full range and to sweep the frequency over a ratio of 1000:1 on any selected range.

The Jupiter 500 comes complete with mains lead and instruction manual and costs £110.00 plus VAT.

Black Star Ltd, 4 Stephenson Road, St. Ives, Huntingdon, Cambridgeshire PE17 4WJ, tel (0480) 62440.

World's Smallest DPM



Datel claim their DM-500 is the smallest digital panel meter in the world. It measures just 48 x 24 x 43mm (1.89 x 0.94 x 1.69") and at around 60mA its power consumption is said to be three times lower than that of comparable DPMs.

The DM-500 has a 3½ digit red LED display with 8mm (0.32") high characters which are designed to be clearly visible even in bright conditions. Different versions provide four DC voltage ranges from ±199.9mV to ±199.9V and four current ranges from ±199.9uA to ±199.9mA. Display accuracy is ±0.1% of reading ±1 digit and each range is provided with over-range signal protection.

The meter samples the incoming signal 2.5 times per second and zeroes automatically. A hold facility allows the sampling to be stopped so that the last signal measured is displayed. Both single-ended and differential signals are accepted and an input impedance of around 1000M ensures that sensitive inputs aren't loaded.

Datel UK Ltd, Belgrave House, Basing View, Basingstoke, Hampshire RG21 2YS, tel (2056) 469 085.

PC-Compatibility For The BBC

A new interface board from Watford Electronics allows BBC B and B+ microcomputers to run PC-type software packages using the Acorn Master 512K board.

As well as opening up new possibilities for the BBC as a serious business machine, the interface makes it possible to link-up a Master Turbo co-processor board and so gain extra speed and memory.

Known as the Co-Pro Adaptor, the interface will cos £50.00 and its launch follows a recent price-cut on the Master 512K board to £199.00. This makes it possible to upgrade a BBC for around £250.00, far less than the purchase price of even the cheapest PC clones.

Watford point out that there are still over 500,000 BBC microcomputers aroudn and claim that the introduction of the new board demonstrates their continuing commitment to the machine. They expect the Co-Pro Adaptor to be popular with schools since almost half of the cost can be claimed back under the DTIs Software In Schools Scheme.

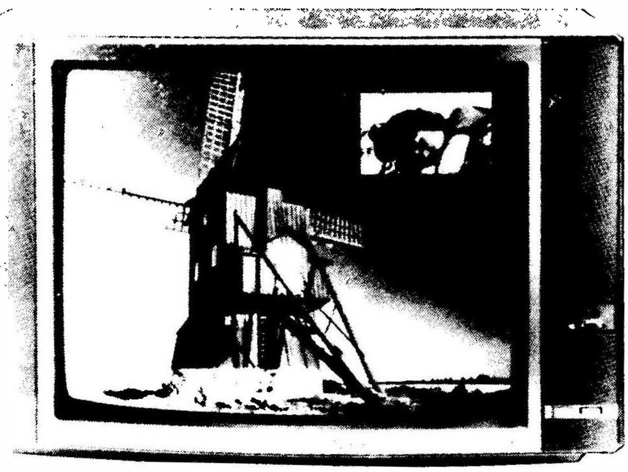
Watford Electronics, 250 High Street, Watford WD1 2AN, tel (0923) 37774.

IC A Picture In A Picture

A new IC from ITT allows a second, one-third size picture to be displayed within the main picture on a TV screen.

The second picture can be placed in any one of four predetermined positions on the screen and is contained within a border which separates it from the main picture.

The IC is part of ITTs Digit 2000 system and is known as the PIP 2250. It works by storing the Y, R-Y and B-Y signals in a DRAM and presenting them to the video controller at the appropriate times. The additional RGB inputs to the controller are used to produce the border and an internal switch allows the PIP2500 to accept external RGB inputs so that Teletext and other picture data can still be used. In keeping with its size, the small picture is sampled at a rate of every third pixel, every third line.



A number of other new ICs have recently been released by ITT, among them a programmable speedometer and mileage indicator system for use in cars and a real-time digital signal processor for digitising audio signals.

The speedometer and mileage indicator consists of two ICs, the SAF1091 and SAF1092, and works by sensing the speed of rotation of the gearbox driveshaft. An LC oscillator drives a coil placed near the gearbox, and a proximity switch within the bipolar SAF1091 senses drop-outs in the oscillation caused by a small piece of metal attached to the driveshaft. The resulting pulse train is fed to the CMOS SAF1092 for logic processing and then returned to the SAF1091 which drives a moving-coil meter to indicate speed and a stepper motor to register the miles covered. The ICs are suitable for both 12 and 24V electrical systems and are supplied in 18-pin DIL packages.

The digital signal processor IC, the APU 2470, is intended principally for filtering operations which it performs by combining add, multiply and delay functions. It has a sampling rate of 35kHz and handles several calculations and move operations simultaneously, achieving a maximum of 4.4 million sum-of-product operations each second. A mask-programmed ROM holds the program and a 51 x 16 bit RAM stores intermediate results. Among the other functional blocks within the IC are a fast ALU, digital filters for input PCM conversion, an output DAC and four 16-bit serial I/O channels. Analogue volume adjustment is provided to ensure maximum signal-to-noise ratio. The APU 2470 is fabricated in N-channel MOS and comes in a 24-pin DIL package.

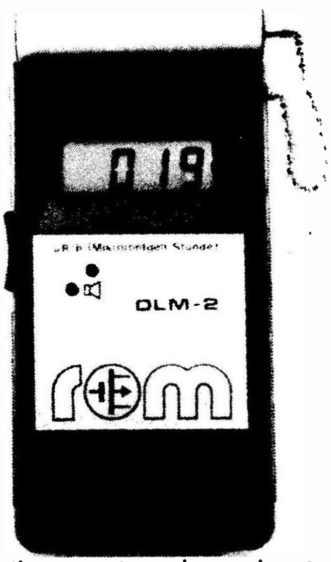
ITT Semiconductors, 145-147 Ewell Road, Surbiton, Surrey KT6 6AW.

Pocket Ratemeter

A little something for those who can't be bothered to build this month's Geiger ratemeter project. From Germany comes the DLM-2, a pocket-sized ratemeter complete with GM tube which responds to alpha, beta, gamma and X-rays. It has an LED and a loudspeaker to indicate the detection of radioactive particles and the manufacturers say it is useful for checking food and drink as well as other items such as clothing.

The measurement range is from 1-1999 uR/h (micro-Roentgens per hour) and the result is displayed on a 4-digit, 0.5" high LCD. The current consumption is said to be very low in order to prolong battery life and a low-battery warning is incorporated in the display. The circuit diagram shows a pulse output socket which could be used to drive an external counter, but no mention of this is made in the notes sent to us.

The manufacturers say they are happy to supply the ratemeter to UK customers but do not mention a postage charge. Anyone interested in buying one is advised to check this carefully before ordering. The cost of the DLM-2 is DM750 which works out to about £214 at the current rate of exchange.



Rober May Elektronische Gerate, Babenhauser Str.55, D-8908 Krumbach 1, Federal Republic of Germany.

High Capacity All Wound Up

ERA Technology has developed a type of capacitor construction which is capable of delivering 50 farads from a package the size of a cotton reel.

Using a wound double-layer arrangement the company has achieved an energy storage density of six joules per cubic metre at low DC voltages. A prototype capacitor has been shown to be capable of retaining its charge over many months and is said to have stable discharge characteristics. Electric double-layer capacitors are already available in capacitances of a few farads for use in computer memory back-up, and ERA suggests that high value capacitors might be used to allow processing as well as memory to be sustained during power down.

ERA would be interested to hear from other companies involved in the development of large value capacitors for use at low DC volt-ages. Contact them at Cleve Road, Leatherhead, Surrey KT22 7SA, tel (0372) 374 151.

● **Educational electronics specialist Flight Electronics has issued its latest catalogue of microprocessor, logic and analogue training systems. Packages for the Z80, 8088 and 68K are among those described and the catalogue also lists technical books and a range of training-related equipment such as bench power supplies, breadboards, etc. Flight Electronics Ltd, Flight House, Ascupart Street, Southampton SO1 1LU, tel (0703) 227 721.**

● *Piezo Products can manufacture small quantities of crystals to a customers' requirements within three working days. The service covers devices in the 1-70MHz range and in most standard packages and offers an accuracy of ±0.001% at normal temperatures. Piezo Products Ltd, Mill-Stream Trading Estate, Christchurch Road, Ringwood, Hampshire BH24 3SD, tel (0425) 479 337.*

● **A new 40-page A4 catalogue from P. Caro & Associates lists over 170 different slide switches with illustrations and full technical information. Ratings from 0.3 to 3 amps are covered with from one to eight poles and up to ten positions. There are also a small number of toggle, rocker and pushbutton switches with higher current ratings and a few jack-type DC power connectors. P. Caro & Associates Ltd, 2347 Coventry Road, Sheldon, Birmingham B26 3LS, tel (021) 742-1328.**

● Three new brochures from Hitachi describe the company's range of liquid crystal dot-matrix displays. One contains specifications on the complete range of devices while the others describe two new product lines, X-Technology displays and LCD modules with built-in LED backlighting. X-Technology is said to offer better contrast and viewing angle than conventional LCDs and is being applied to large displays which would normally lack contrast because of multiplexing. The LED backlighting system is offered as an alternative to high-voltage electroluminescent systems, having longer life and running from a 5V supply. Contact Hitachi Electronic Components (UK) Ltd, 21 Upton Road, Watford, Hertfordshire WD1 7TP, tel (0923) 46488.

● *Also available from Hitachi is a 32-page A4 data book on the HG62B series of CMOS gate arrays. A list of macrocell configurations is given, plus notes and examples on the design of logic systems and a guide to the use of auto-diagnosis and other features. Contact Hitachi at the address above.*

NEXT MONTH

AN ARGUS SPECIALIST PUBLICATION

electronics today INTERNATIONAL

HI-FI FM TUNER

John Linsley Hood, that last bastion of quality in the British hi-fi market, brings you his latest design in the pages of ETI. JLH's tuner is based on novel principles but those of you who have followed his recent series on tuner design will know to expect the best.

THE ETI CAPACITOMETER

From a few p's to lots of μ's the ETI Capacitometer can measure the exact value of any capacitor. It's portable and quick to give a reading — a valuable addition to any workshop.

COMPUTER DATA COMMUNICATIONS

Baffled by busses? Perplexed by parallel? At sixes and sevens with serial? All is explained in next month's low down on the world of computer interfacing standards.

AND THE REST

All our regular features will be there too. Projects, features, letters, news, reviews, readers' ads and much much more.

THE MARCH ISSUE OF ETI — ON SALE 6th FEBRUARY

YOU'D BE MAD NOT TO GET IT!

All the articles above are at an advanced state of preparation, but circumstances beyond our control may prevent publication.

Standing Up For More Memory

Texas Instruments has found a simple way to get four times as much memory into one IC. Instead of laying chips horizontally within the package, they are stood on end. That way, far more circuitry can be contained in the same amount of space.

The new storage concept is being employed in the 4MB DRAM currently under development. One of the advantages of the method is that it achieves higher density without the need for new manufacturing tools. TI claims that only one other manufacturer has been able to produce a 4MB DRAM of similar size and a complete new generation of production equipment was needed to do it.

Texas Instruments Ltd, Manton Lane, Bedford MK41 7PA, tel (0234) 63211.

Never A Crossword

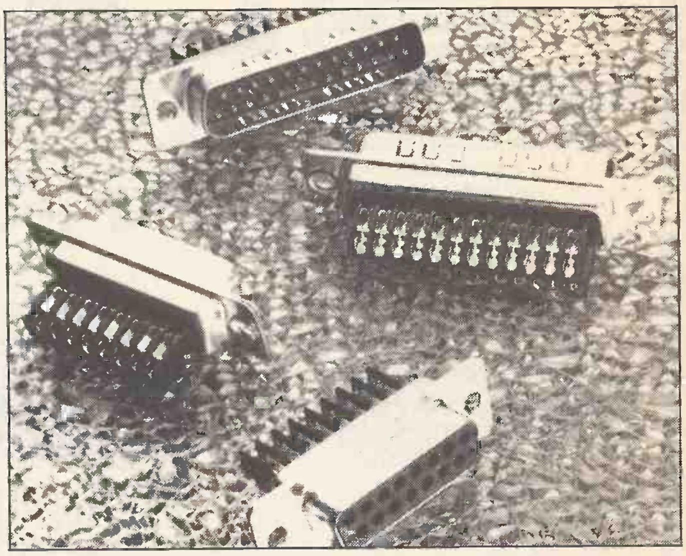
Devoted fans of the ETI crossword will be disappointed to learn that it won't be published this month.

We started publishing the crossword just over a year ago, and since then it seems to have drawn praise and criticism in roughly equal measure. We have now used all of the original twelve crosswords we commissioned but we hope to begin a further series next month. If you have any strong feelings about what we should do with the crossword, why not send a letter in to Read/Write?

Meanwhile, here are the answers to last month's crossword.

ACROSS
1) Throat
7) Triac
8) Algebra
9) Switch
11) Tesla
12) DIN
14) Atom
15) Address
16) Read
17) Bel
18) Bench
19) Stator
22) Flyback
23) Spark
24) Tweeter

DOWN
2) Hall effect
3) Overload
4) Turn
5) Brown
6) Mast
9) Siren
10) Clock cycle
13) Adder
14) Assemble
17) Board
20) Amps
21) Blow



Good As Gold

The latest range of IDC connectors from Dage uses a contact plating system which is said to be better than gold but which costs far less.

Known as Robex, the plating employs a thin layer of gold to ensure low contact resistance but underlays it with nickel and nickel-palladium. The result is that far less gold is required for a given plating thickness. Because the Robex plating covers surface imperfections better than gold does, the material is less porous and has greater resistance to corrosion, wear and soldering. Dage claim that a ten microinch plating of Robex will perform as well as 30 microinches of gold in standard corrosion and porosity tests and offers virtually the same contact resistance.

Dage (GB) Ltd, Rabans Lane, Aylesbury, Buckinghamshire HP19 3RG, tel (0296) 35408.

DIARY

Television Engineering Course — January 7th

The IBA, London. First of fifteen weekly lectures. Contact the Royal Television Society at the address below.

Satellite Navigation Systems — January 8th

The IEE, London, 5.30 pm. The 22nd Appleton Lecture, given this year by Dr. Peter Daly of the University of Leeds. Contact the IEE at the address below.

Digital Techniques In Television — January 13th

RTS Southern Centre, Newbury. First of eleven weekly lectures. Contact the Royal Television Society at the address below.

The Impact of ASIC on Equipment Design — January 14th

The IEE, London, 6.30pm. Colloquium on Application Specific Integrated Circuits. Contact the IEE at the address below.

Design and Measurement for EMC/RFI Management — January 19th

The IEE, London, 2.00pm. Discussion meeting. Contact the IEE at the address below.

Television Studios from Analogue to Digital — January 21st

The IEE, London, 6.30pm. Contact the IEE at the address below.

The Third High Technology and Equipment in Education Exhibition — January 21-24th

Barbican Centre, London, Exhibition and conference covering all areas of educational and training technology, etc. Contact Christine Smith on 01-608 1161.

Testing and Inspection of Electronic Components and Circuits — January 22nd

The IEE, London, 6.30pm. Contact the IEE at the address below.

Digital Signal Processing — January 26th

The IEE, London, 6.30pm. Contact the IEE at the address below.

SMARTEX '87 — February 10-12th

The Barbican Centre, London. The 2nd Surface Mount and Related Technologies Exhibition. Contact Peter Evans on (01) 855-7777.

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Kensington Rainbow Exhibition Centre, London. Exhibition and conference devoted to power sources and supplies, converters, etc and the associated areas of enclosure screening, thermal management etc. Contact TCM Expositions on (0428) 724 660.

The Electronic Data Interchange Conference — April 28-29th

The Barbican Centre, London. Conference which aims to explore the possibilities for improved electronic communcation within and between organisations nationally and internationally. Contact Nicky Cross of Online on (01) 868-4466.

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READ/WRITE

Why MIDI

Now that you are covering the subject I've just got to ask. Why MIDI? I can see that it is awfully clever to send data between musical instruments at around 30K a second, but why would I want to? I should admit I am no musician, but it seems little short of a waste of time to connect up assorted synthesisers and drum machines as indicated in your feature in the January issue.

Surely, if you need a different sound or require lots of noises at once, the answer is to get a different instrument or more musicians, not to connect up all sorts of keyboards together.

I realise I must be wrong because everyone is out there buying these things, but why?

George Beecham.
Morecambe, Lancs.

It would be nice if we could all afford all the instruments (and all the instrumentalists) we ever needed. Instead, MIDI provides a way for a single musician to perform with the help of a computer with all the benefits its programmability allows. The computer can be a micro or just the innards of a synthesiser or a drum machine. Either way, the computer can be programmed to interpret the player's actions as required for that particular song. The interconnectability of MIDI allows, say, half of a keyboard to 'play' one instrument while the other half plays an altogether separate one. This is at the very least much easier than the old Rick Wakeman-esque method of stretching across a great array of keyboards.

While on the subject of MIDI, last month's feature showed a photograph which we incorrectly captioned as showing Hybrid Technology's Music 5000 system. In fact it was a UMI-2B MIDI controller for the BBC Micro. Apologies to both parties for the mix up. Readers interested in music on the BBC Micro will find plenty to interest them in the coming months in ETI.

Collector's Corner

I am a collector of electronic calculators. I am also trying to trace the development of these now common devices since they first began to appear circa 1970. To this end I have been ploughing through many old copies of ETI, which contain a wealth of information in articles and adverts.

It occurred to me that many readers might have info I would find useful, possibly in the form of old instruction booklets and the like, for calculators that have long since 'bitten the dust'.

If anyone has any such info they would care to donate to posterity, I would be pleased to hear from them.

Bryan Cloud,
Ipswich, Suffolk.

If any readers have old calculator manuals or even old calculators themselves, they can contact Mr. Cloud on (0473) 44996.

LP RIP

I wonder if other readers regard the advent of Compact Disc with the same concern as I do.

Although I would not deny the technological advances involved, there appears to be no provision for people like myself with large collections of conventional vinyl records.

It is unlikely the majority of these will ever appear as Compact Discs. As conventional record decks and other equipment are taken out of production the supply of replacement parts and spares will dry up.

Personally, I hope the dominance of Compact Disc over the rest of the hi-fi market will not occur, at least until a method of recording Compact Discs at home has been perfected.

Paul Berry,
Acrrington, Lancs.

It is true that many of your records will never be issued as Compact Discs (although who would have thought old black and white, and even silent, movies would be commercial candidates for transfer to video — but they are). However, the vinyl record has a good few years left in it yet. Millions of people have 'conventional' record decks and many more millions of records are sold each year. No record company is going to throw away that kind of revenue if it can help it.

There *will* come a time when vinyl records are no longer produced. However, at that stage people will no doubt be bemoaning the advent of the next leap forward which will make their CD players redundant. Meanwhile, digital cassette systems are only around the corner (no, really!). These will give a quality on a par with CD *and* allow you to record your old collection.

You cannot stand in the way of progress. CDs will come and eventually go. Some consumers will be left high and dry without spares or musical media but in the long run (as governments say) we will all be better off.

Picture This

Although I am sure you can do without a constant stream of suggestions' as to how to improve ETI (excellent as it is already) I feel I must complain about the January issue.

It is not the projects I am winging about. These were as interesting as ever. The features were also up to their usual informative standards. No, what I want to suggest is that you improve the quality of the photographs in ETI. In the January issue they were quite honestly appalling. It looked like there was a heavy fog settling in the photographer's that day!

I won't ask you to pull your socks up. Perhaps they are already pulled over the camera lens!

Malcolm Walmsley,
Sheffield, South Yorkshire.

To be fair to our photographers the fault with the January issue was in the production process of the magazine and not under the stuido lights. However, you are quite right in your assesment of the quality of reproduction. We have taken steps and hope that this issue fares better. Incidentally, we always appreciate suggestions (or even 'suggestions') as to how to improve ETI.

Free Praise

I advertised my Hitachi single beam oscilloscope in a Readers' Free Ad in the November issue of ETI. I received and accepted an offer of £47 (I'd put it in for £50) within seven days. Other enquiries continued to come in.

Many thanks for the free advert.

A. Wallwork,
Sutton Coldfield, West Midlands.

The response to the Free Readers' Ads has been a little disappointing. Although by all accounts those readers who have advertised have found suitable buyers, we are not receiving enough advertisements. Don't let your old equipment rust in a corner. Turn it into cash. Post the coupon today!

ETI welcomes letters from readers on any topic. If you disagree with our learned contributors or just think the world is going to pieces, don't just sit there. Tell us all your troubles. We can't promise to solve anything, but we can print your letter.
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A QUESTION OF BALANCE

Barry Porter takes a fresh look at a frequently misunderstood topic — the use of balanced lines for audio signal connections.

To most hi-fi enthusiasts, a 'balanced' system is one which delivers approximately equal amounts of Max Bygraves from each channel while 'unbalance' is a condition brought about by Junior poking an exploratory finger through one of the loudspeaker grilles. It's well known that something called balancing is also important in the world of professional audio, but not many people seem to know what the system does or what advantages it brings. If you fall into this category and would like to know a little more, read on.

The individual units of a domestic system — such as a preamplifier and a power amplifier — are usually connected together with a length of screened cable (Fig. 1). Signal is carried along the inner conductor, and the outer screen joins the earth lines of the electronics in the two items of equipment. This type of interconnection is termed 'unbalanced' and works well enough providing the cable is not of excessive length and does not pass near to a strong source of electrical interference. Although the screen layer gives some protection to the signal carrying conductor, a strong external AC field such as that surrounding a mains transformer will pass through the screen and add its unwanted modulation to the signal.

This problem is obviously at its greatest when the cable carries signals at a low level and is followed by lots of amplification. An example is the connection between a microphone and the input channel of a mixing console, where any hum and noise picked up by the cable may be amplified by as much as 60dB.

One way to minimize the effect of external hum and noise is to keep the line impedance as low as possible and the signal level as high as possible. If the sending amplifier has a very low output impedance — less than 1 ohm is possible with operational amplifiers — the connecting cable appears to be at about earth potential, and external fields will exert very little influence.

This principle works well for cable runs of less than 30 metres or so where the external field is not too intense, but when long interconnections are used, the resistance of the signal carrying wire becomes significant and the line impedance increases as you get further from the sending amplifier. In some cases, it is not possible to achieve a near-zero source impedance.

A typical moving coil microphone has an impedance of 200 ohms, and some equipment which has passive attenuation or equalisation after the final stage may have an output impedance of several kilohms — ideal circumstances for the introduction of noise whenever connected to a few metres of cable.

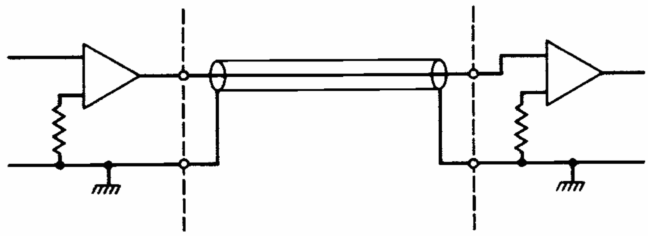


Fig. 1 An unbalanced signal connection, as used on most domestic audio equipment.

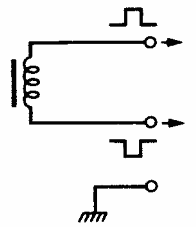


Fig. 2 A balanced output as provided by a moving coil microphone or a transformer.

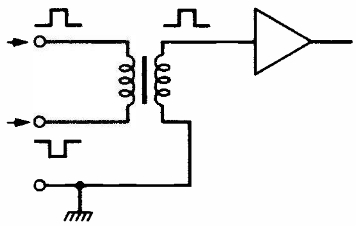


Fig. 3 A differential (or balanced) input using a transformer.

Yet microphones in particular are often required to be at the far end of several hundred meters of conductor, along which a signal with a level of 1mV or less must travel with as little corruption as possible. Can anything be done to minimize the pick-up of hum, noise and Radio 1 that doesn't require considerble lengths of Mu-metal conduit piping? Luckily, it can, and it's called balancing.

The easiest way to understand how balanced connections work is to consider a moving coil microphone. It will consist of a thin foil diaphragm to which is attached a lightweight coil, suspended in a powerful magnetic field. The diaphragm and the coil are vibrated by incoming sound waves, and this vibration is turned into an electrical current by the action of the magnet. The ends of the coil are connected to the microphone output socket (Fig. 2) where the output signal voltage appears with a phase difference of 180 degrees between the two contacts.

For a few minutes, let us turn to a type of input circuit usually referred to as differential and typified by the transformer arrangement shown in Fig. 3. In order to produce an output voltage, the transformer inputs must receive two signals which are identical in every way except that they must exhibit a 180 degree phase difference — just what the microphone gives out.

If the two outputs of the microphone are now connected to the two transformer inputs, a balanced circuit is created. In practice, the two conductors are enclosed by a screen layer which in the example given would connect the metal case of the microphone to the input amplifier earth. This is shown in Fig. 4 along with the amplification stage which follows the input transformer.

You may wonder what is so different between that and an unbalanced connection. Luckily, there is a big difference. If the cable connecting a balanced output to a balanced input picks up any hum and noise, the unwanted voltages will occur in both conductors at the same level, and most importantly, with the same phase. Since a balanced circuit requires a 180 degree phase difference between its two inputs, any signals arriving which are not in anti-phase are cancelled by the differential action of the input. This removes the hum and noise.

The ability of a balanced input to reject unwanted signals is called the Common Mode Rejection Ratio (CMRR for short), and is measured as shown in Fig. 5. CMRR values tend to be frequency conscious, rejection becoming progressively worse with increasing frequency. By careful design, particularly with solid state inputs, a rejection of 80 or 90dB across the audio band may be achieved.

Transformer Or Solid State?

Until recent years, the standard balanced circuit invariably employed transformers for both input and output coupling. By using transformers the signal can be made to pass from the output of one piece of equipment to the input of another without an earth reference, so removing the risk of hum loops. This is particularly important in radio and TV studios, so transformer balancing is almost universal in these environments.

Transformers do suffer from a number of shortcomings. Without becoming too large for comfort, they will only pass a limited signal level before introducing large amounts of very nasty distortion, particularly at low frequencies. They can also suffer from undesirable phase shifts unless carefully designed. High quality transformers are also rather expensive — 'Professional Grade' versions can cost as much as £50–£60.

Although balancing (and unbalancing) transformers are still used in large quantities, for many applications they are being rapidly replaced with solid state input and output stages. A standard operational amplifier has a differential input, and can be connected to a balanced line, as shown in Fig. 6. To obtain the best common mode rejection, the four resistors must all be of the same value, and it is usual to use a variable and fixed resistor in series for R3 to allow the CMRR to be trimmed at low frequencies. The two capacitors also need to have the same value, and C2 is often a small trimmer in parallel with a fixed component that is lower in value than C1. With careful adjustment, the CMRR can be as high in value as 90dB across the audio band. This circuit can give excellent results, but has a few shortcomings which rule out its use in high quality equipment.

The two inputs have differing impedances because R2 is terminated by a virtual earth whereas R1 is loaded by R3 in parallel with the input impedance of the op-amp. Unequal input impedances may not be of too much concern depending upon the type of balanced output connected to them, but the actual impedance value may assume a greater importance. It is standard practice to make the impedance of line level inputs 10k ohms or greater so that signal level is not reduced by loading effects, and it is often desirable to present a load of 50k ohms or more to the incoming signal. It will be found that if the value of the resistors is increased sufficiently to give a reasonably high input impedance — say 33k — then thermal noise has to be considered.

The noise voltage of a resistor is given by

$$\sqrt{4kTBR}$$

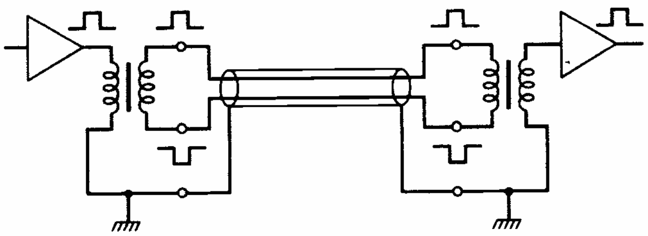


Fig. 4 A balanced interconnection between two pieces of equipment.

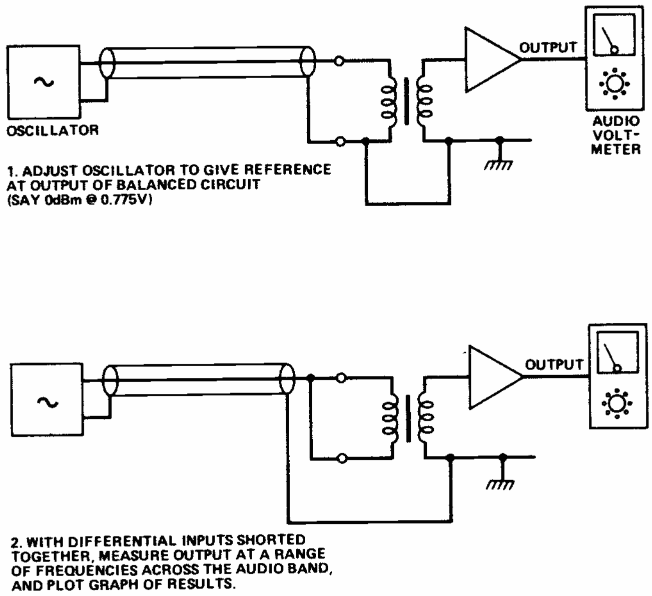


Fig. 5 Measuring the Common Mode Rejection Ratio (CMRR) of a differential input.

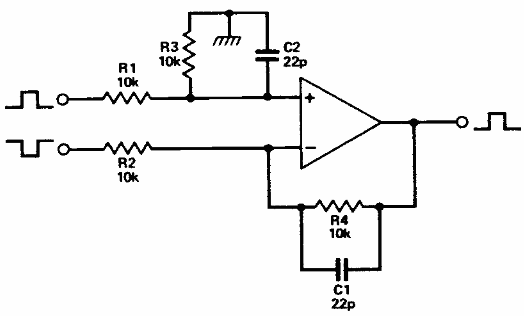


Fig. 6 Balanced input circuit using a single op-amp.

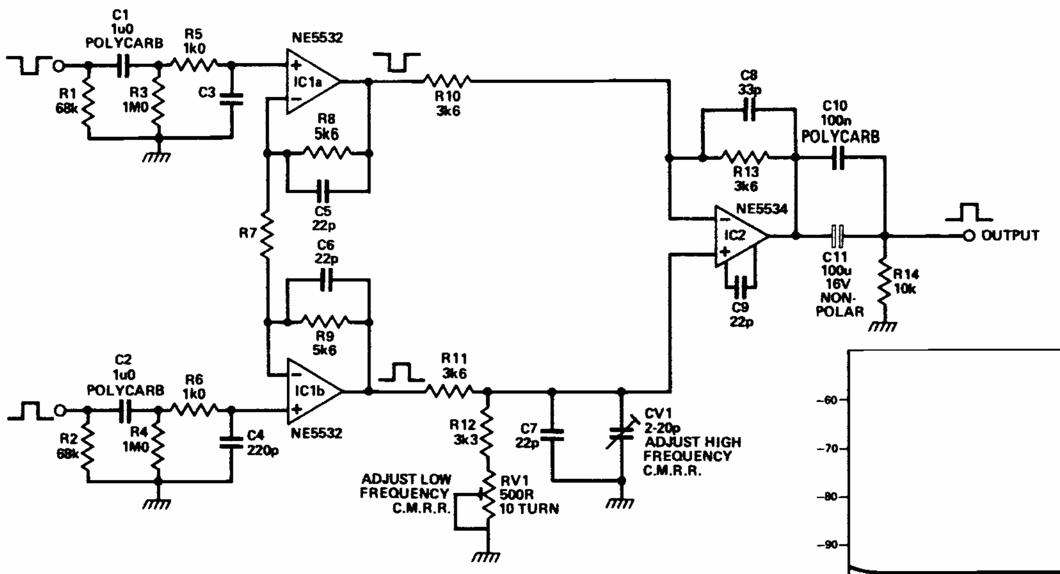


Fig. 7 A balanced input amplifier which uses input buffers to reduce the values of resistance required, thereby achieving a low level of thermal noise.

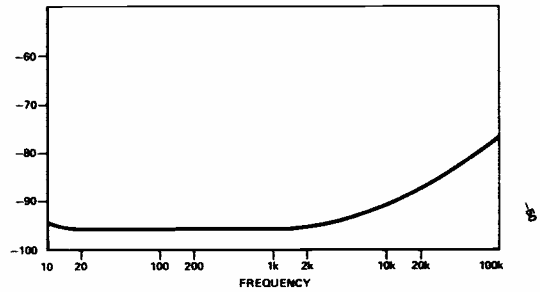


Fig. 8 The common mode rejection offered by the circuit of Fig. 7.

where k = Boltzmann's constant (1.381×10^{-6})
T = Absolute temperature (°C + 273)
B = Bandwidth
R = Resistance in ohms

For a 33k ohms resistor, assuming a band width of 20kHz and an ambient temperature of 25°C, the noise voltage is therefore 3.296×10^{-6} or, in a more recognizable form

$$20 \log \frac{0.775}{(3.296 \times 10^{-6})} = -107.4\text{dBm}$$

which is not particularly brilliant. Now, if the resistor value is reduced by a factor of 10, its noise contribution will be reduced by 10dB, which is a significant improvement.

A practical circuit which takes account of the thermal noise problem is shown in Fig. 7. It is essentially the same as the circuit shown in Fig. 6 except that buffers have been added to each of the inputs. These allow the impedance of each input to be the same and of virtually any value required, yet the resistances around the differential stage may be low enough to avoid noise problems. An additional advantage is that the stage gain can be adjusted by changing one resistor — R7. The gain is calculated from

$$20 \log \left(\left(\frac{R8 + R9}{R7} \right) + 1 \right)$$

or for a required gain,

$$R7 = \frac{R8 + R9}{A \log \left(\frac{Av(dB)}{20} \right) - 1}$$

This stage is ideal for line level inputs operating over a gain range from unity to 20 or 30dB. It has low noise, negligible distortion (typically 0.002% over the audio band) and excellent common mode rejection (see Fig. 8). To adjust RV1 and CV1, use the set-up shown in Fig. 5. With a 100Hz input signal, set RV1 so that the circuit output voltage is at its minimum level, then reset the oscillator to 15kHz and trim CV1 for minimum output.

This circuit is often referred to as an 'Instrumentation' input. As well as being an ideal balanced input stage for pre and power amplifiers, mixing consoles etc, it also makes a suitable input for many types of high quality test equipment. Audio frequency voltmeters, distortion meters and many other instruments use the circuit because it has very stable and clearly defined gain characteristics and a wide bandwidth with an extremely flat frequency response. With the component values shown in Fig. 7, the -3dB points are at approximately 0.5Hz and 150kHz with less than 2° of phase lead at 20Hz and about 7° of lag at 20kHz.

A number of changes may be introduced to optimize the circuit for specific applications. Input noise may be reduced by replacing C1 and C2 with 10μ or 22μ non-polarized electrolytics (not forgetting to put a 100n polycarbonate in parallel with each one in the interest of sound quality!) This will also give a few more dB of common mode rejection at low frequencies, and the phase shift at 20Hz will be reduced to 0.06. For use in test equipment, C3 and C4 should be reduced to 47p in order to keep the response flat to 100kHz.

Stage gain may be made adjustable by using a single pole switch to select different values of R7 or a potentionmeter in series with a fixed resistor (see Fig. 9). Note that AC coupling has been introduced to avoid problems from switch clicks or potentiometer wiper noise which would otherwise arise due to inevitable DC offsets at the op-amp inputs.

If an accurate stepped attenuator is placed in front of this circuit and some additional gain stages added after it, you have the input stage of a professional-quality audio voltmeter which will give FSD readings from 300V to 10uV — a range of 150dB! (When some further development work has been carried out, you may have the pleasure of a constructional article based on it).

So far, we have looked at the balancing of high level signals, but there are far more benefits to be gained by employing the same techniques when signal levels are in the milli or microvolt region. For a high quality microphone amplifier, the basic configuration of Fig. 7 may be used but with a few important changes.

The biggest single problem with microphone amplifiers is keeping the noise contributed by the active stage as low as possible. The thermal noise of a 200 ohm source is –129.6dBm, so a perfectly silent microphone amplifier with a gain of 60dB and a 200 ohm input load, would have an output voltage of –69.6dBm over a 20kHz bandwidth. If a real amplifier has an output of –67dBm, it is said to have a noise figure of 2.6dB. Some manufacturers claim noise figures as low as 1dB, but 1.5dB is a more realistic aim, and the circuit of Fig. 10 can achieve this providing great care is taken with construction and component choice.

The input impedance is fixed at 1k2 ohms by R4, this being the currently accepted value for loading 200 ohm microphones. It may be changed if a different value is recommended by the microphone manufacturer. R1 and R2 apply 48V phantom power to energise capacitor microphones and should be closely matched, either by purchasing 1% tolerance components or by measuring with an accurate bridge or digital multimeter. If the circuit is only to be used with dynamic microphones, R1, R2 and R3 together with C1 may be omitted, as may C2 and C3. This will give a slight improvement in signal-to-noise ratio and an improvement in sound quality due to the loss of a pair of electrolytic capacitors from the signal path. As the common mode rejection is adjusted at a part of the circuit that comes after the gain setting stage, there will be a variation of rejection with gain.

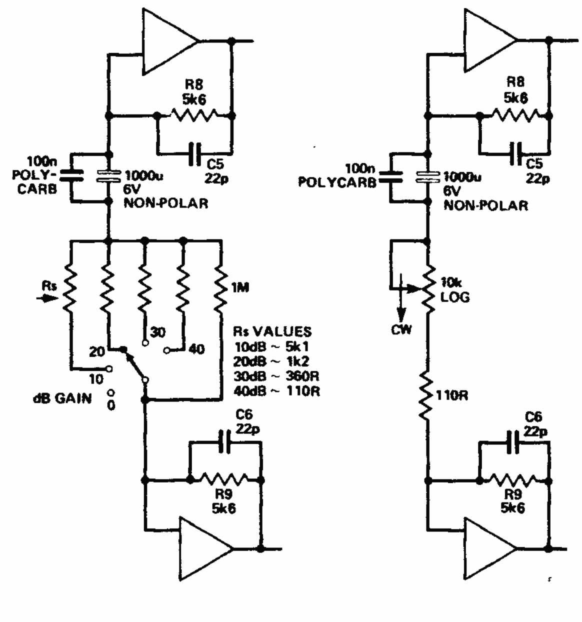
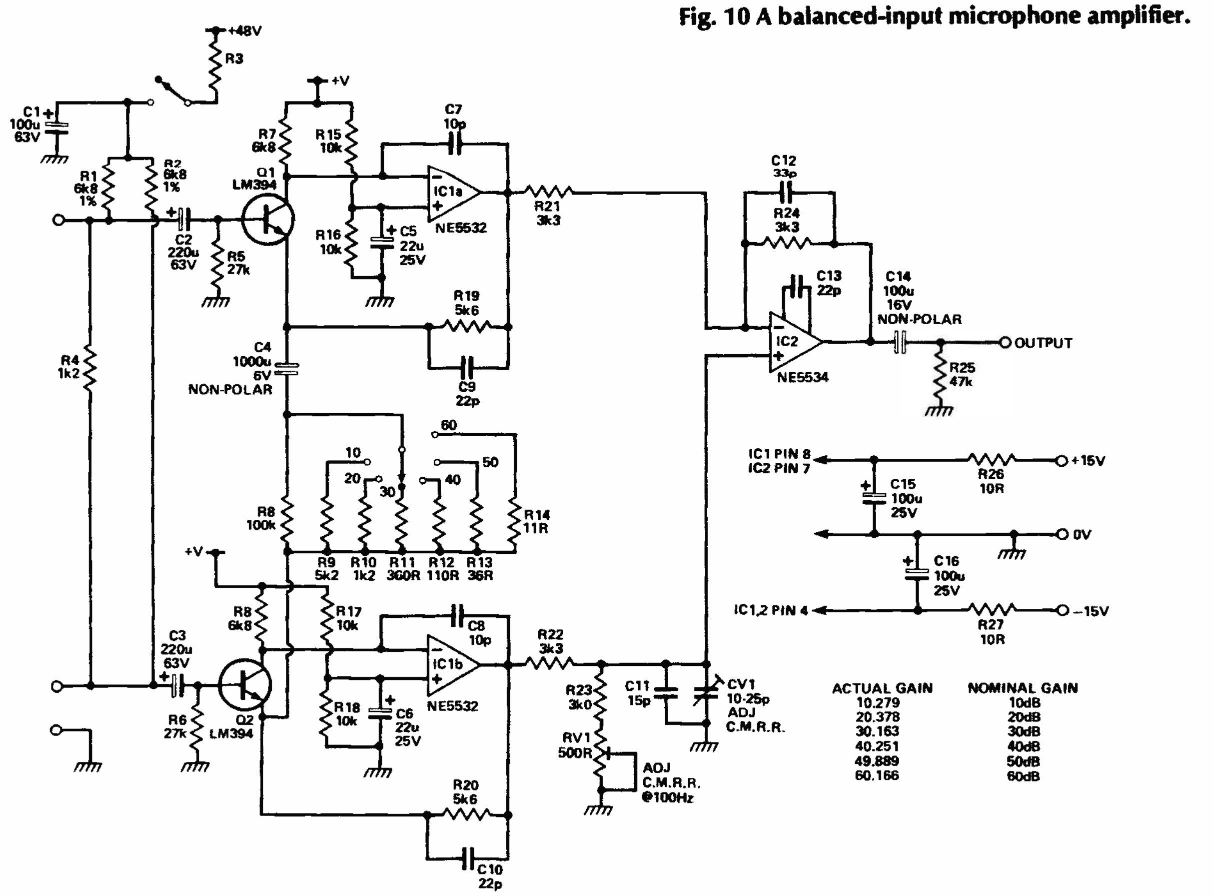


Fig. 9 The circuit of Fig. 7 can be modified to provide either (a) switched gain or (b) continuously variable gain.

Although more impressive rejection can be obtained at 10dB gain than at 60dB it is preferable to carry out the adjustment at the 30dB or 40dB setting.

With a carefully designed PCB layout, this circuit will perform as well as and possibly better than any professional microphone amplifier currently available.

Fig. 10 A balanced-input microphone amplifier.



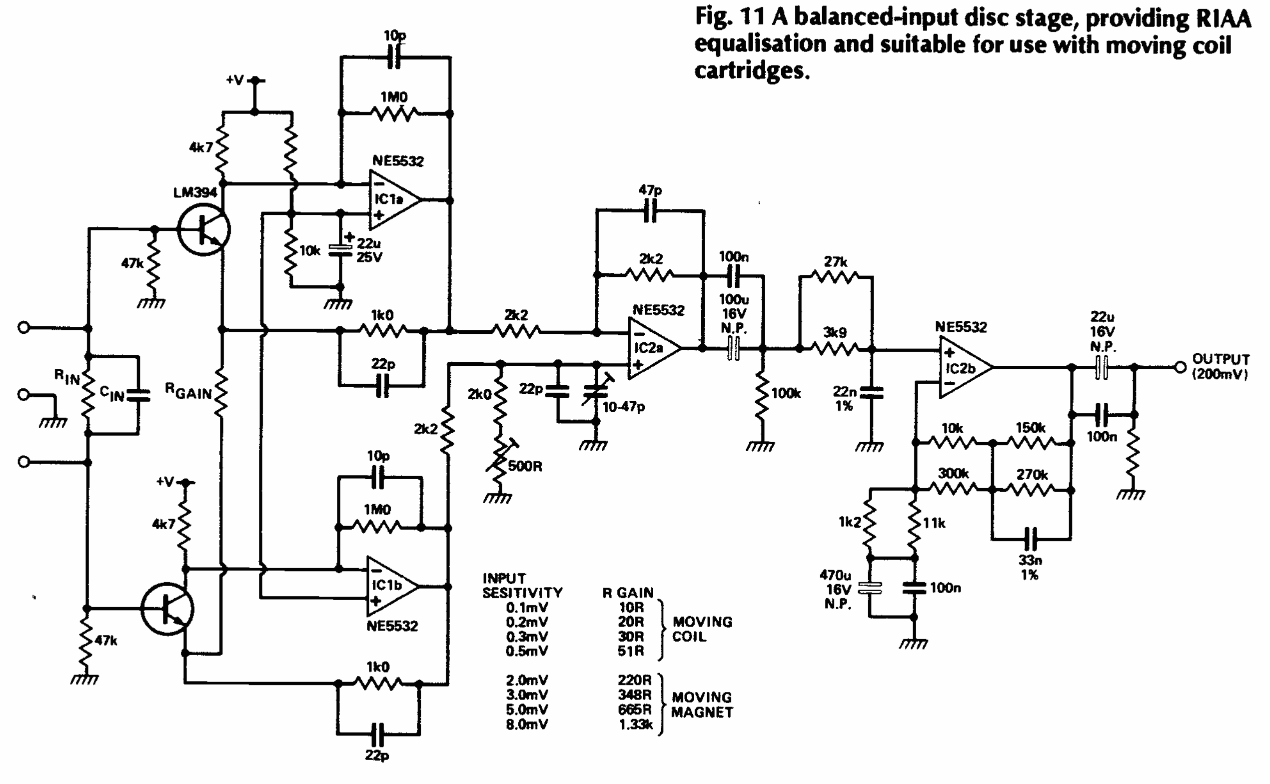


Fig. 11 A balanced-input disc stage, providing RIAA equalisation and suitable for use with moving coil cartridges.

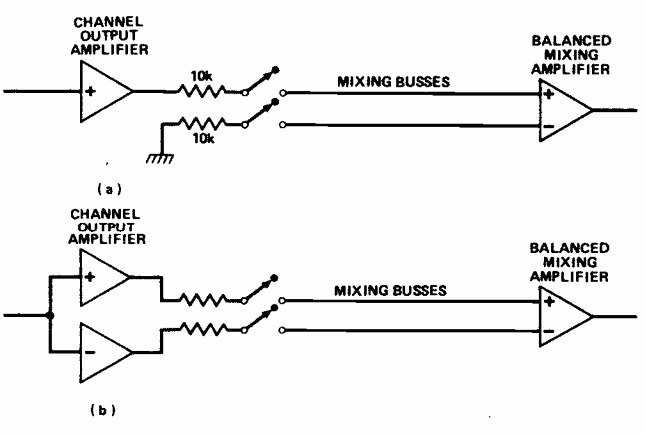


Fig. 12 Balanced mixer arrangements using (a) single-ended amplifiers and (b) balanced-output channel amplifiers.

Balanced Disc Stage

Another type of input stage that benefits from balancing is the RIAA pick-up amplifier, particularly when a low output moving coil cartridge is in use. Figure 11 shows the disc input of the Mirage C10 control Unit, which has a balanced input. Using such an arrangement may mean that minor surgery has to be carried out on the wiring of your pick-up arm in order to remove the ground connections from the signal leads, but the results will be found to justify this annoyance. Try it and see.

Balanced Mixing

The final balanced input circuit I have shown is that of a virtual earth mixing amplifier. Mixing busses are extremely sensitive to hum and noise, but the problem can be greatly reduced by adopting a balanced mixing system. Two basic arrangements are possible, the input channel amplifiers having either single-ended outputs or balanced outputs (Fig. 12). In both cases the signals feed into a differential input, and any noise voltages picked up along the bus lines will appear at the mixing amplifier inputs with the same phase and will be cancelled in the usual way.

A practical circuit for a differential input mixing amplifier is shown in Fig. 13. It resembles the line input circuit of Fig. 7, except that here the incoming signals are taken to the inverting inputs of the op-amps. The use of 3k3 resistors for feedback and mixing helps keep the noise down and the linearity is improved by having the input capacitors inside the feedback loop. With 3k3 input resistors, this circuit will accept up to fifty inputs with no low frequency problems and will attenuate common mode noise from mixing busses ten or twelve feet long by 60–80dB.

Balanced Outputs

Microphones and pick-up cartridges are intrinsically balanced, but amplifier outputs are not. The easy way out is to hang a transformer onto the end of a single-ended output stage, but in this day and age it is rumoured that miraculous things can be done with a handful of chips and a bit of ingenuity.

If a balanced output is just two outputs in phase opposition, there are several ways of obtaining this (see Fig. 14). They all have one failing — namely that if one output is shorted to ground, the output level drops by 6dB because the peak-to-peak voltage swing is halved.

A solution is shown in Fig. 15, a balanced output stage which delivers the same level into balanced

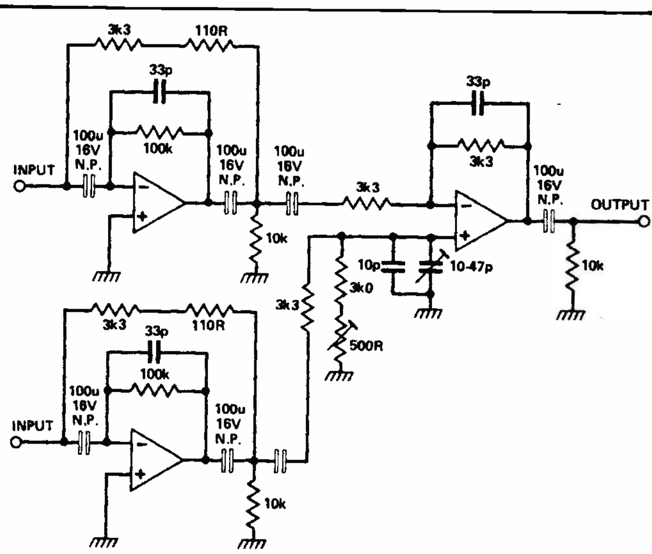


Fig. 13 A practical circuit for a balanced-input mixing amplifier.

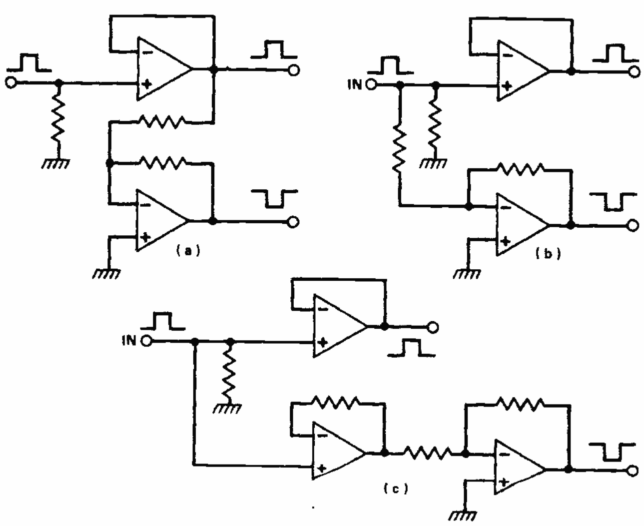


Fig. 14 Some differential (or balanced) output configurations.

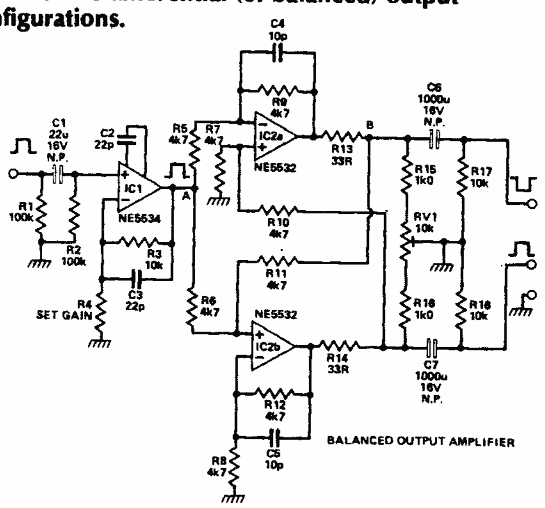


Fig. 15 A balanced output amplifier. The cross-coupled feedback increases the output of either op-amp by 6dBm when the opposite output is shorted to ground, thus allowing the circuit to drive unbalanced inputs as well as balanced ones.

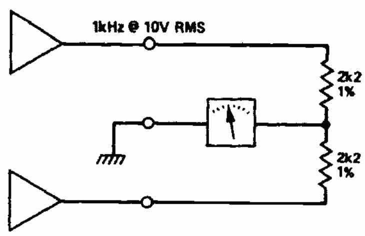


Fig. 16 Adjusting the balance of the circuit shown in Fig. 15. RV1 is set to give a minimum output level.

and unbalanced inputs. When this is used with an unbalanced input, it is essential that the unused output is shorted to ground as distortion and noise will otherwise increase.

The operation of this circuit is fairly obvious. Disregarding the input buffer, assume a 1V input at A which feeds both the inverting and non-inverting op-amps. Taking the inverting stage first (IC2a), you may expect there to be –1V at point B. Instead the cross-coupled feedback through R10 is attenuated by the potential divider R10, R7 and puts +0.25V onto the non-inverting input of IC2a. This is then amplified by 2 (think of R5 as a feedback shunt resistor connected to 0V) so the output becomes $-1+(0.25x2) = -0.5V$. The non-inverting amplifier (IC2b) has a gain of 2 due to R8 and R12. Its input is held midway between the input voltage and the voltage at B due to the dividing action of R6 and R11. The output of IC2b is therefore

$$2\left(\frac{-0.5+1}{2}\right) = +0.5V$$

The output between B and C is consequently 1V.

In the unbalanced mode with output C grounded, the feedback voltage via R10 no longer exists, so IC2a with its gain of 1 gives an output of –1V at B. if B is shorted to ground, the voltage at the non-inverting input of IC2b becomes +0.5 which is amplified to 1V and appears at C.

Unfortunately, this balance is not so easy to achieve in practice, primarily due to component tolerances. To overcome this, a preset (RV1) is used to load one output more than the other until the balance is exact. This should be set so that both outputs are exactly the same level when measured with respect to ground. An easy way to make this adjustment is shown in Fig. 16.

So, now you know how to balance. The question remains, do you know what to balance? Any interconnection between two pieces of audio equipment can be balanced by the addition of a few op-amps, but it is rarely worth interfering with low impedance unbalanced connections which are carrying high level signals over short distances — say less than 20 metres. Microphones should always be fed to balanced inputs, and it is also worth balancing pick-up cartridges and tape recorder heads. Connections between pre and power amplifiers are often improved by balancing, especially where earth currents from the power amplifier would otherwise find their way into the sensitive parts of the pre-amplifier and cause hum.

In brief, balancing can only bring about an improvement in a situation, so if you have more than your share of the dreaded hums give it a try. **ETI**

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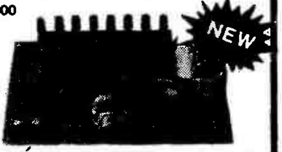

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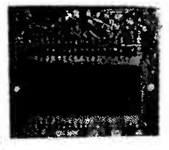

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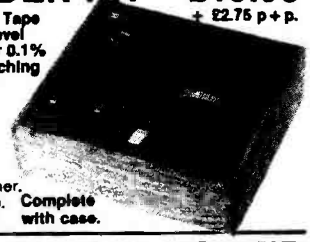

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HARDWARE DESIGN CONCEPTS

Mike Barwise proves his practical approach to hardware design.

Having spent several months discussing theoretical design principles, I would now like to take you through a real project. Rather than present this as a conventional ETI construction project, I will be demonstrating the design process and the step-by-step decision-making culminating in a fully-fledged circuit and PCB. The final design will incorporate both fast logic and an embedded microprocessor.

Test gear is always useful, and for digital and microprocessor work the pulse generator is one of the fundamental tools. Pulse generators tend to be either appallingly expensive or very limited, so I have come up with a DIY solution which is versatile and relatively cheap.

In order of priority, the main requirements of a pulse generator are:

- Reasonably broad range of square wave output (say 100 Hz to 1MHz minimum).
- Provision for delayed short pulse within overall period (Fig. 1)
- Good stability (the ability to remain at the selected settings).
- Reasonably precise setability and repeatability (the possibility of selecting a specific setting or of returning to previous settings after adjustments).
- Auxiliary outputs with shiftable voltage levels.

Most low cost commercial pulse generators use cascaded monostable blocks (Fig. 2a) as the timing chain. With switched capacitors they can easily fulfil the range criterion, and the delayed pulse facility and shiftable levels are standard. However, stability tends to be temperature dependent, and setability and repeatability are poor, normally requiring the use of an oscilloscope to set up critical pulse parameters. This reduces the overall precision to that of the 'scope, (±3% or so).

The Digital Alternative

The approach which we will adopt is to use digital timing chains instead of analogue monostables (Fig. 2b). If properly implemented, the overall precision will be only marginally less than that of the master oscillator which drives the system. The use of a crystal component will provide a system precision in the order of one part in 1000 or better. As the timing chains use digital counters, setability and repeatability are absolute, and the overall range of the generator depends simply on the maximum count of the digital counter chains.

There is a wide variety of counter devices in each conventional logic family. It is important to choose a suitable device for the purpose. I have chosen TTL for our pulse generator, simply because 4000 series CMOS (the major alternative) is too slow, operating at about 2MHz max. As we want our pulse generator to work at up to 10MHz, the choice is clear.

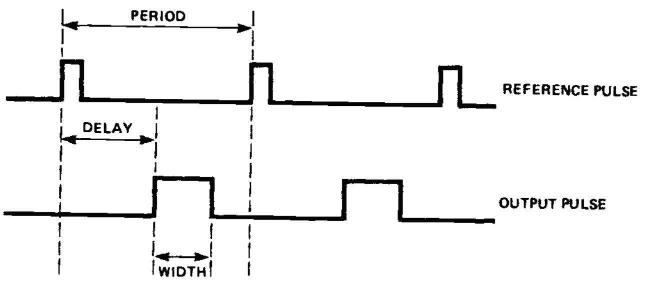


Fig. 1 Pulse delay and width control.

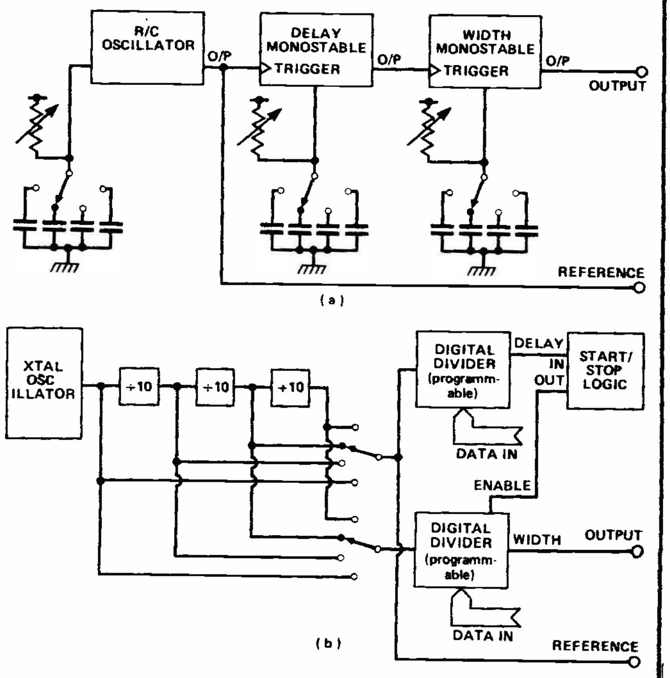


Fig. 2 (a) Analogue pulse generator. (b) Digital pulse generator.

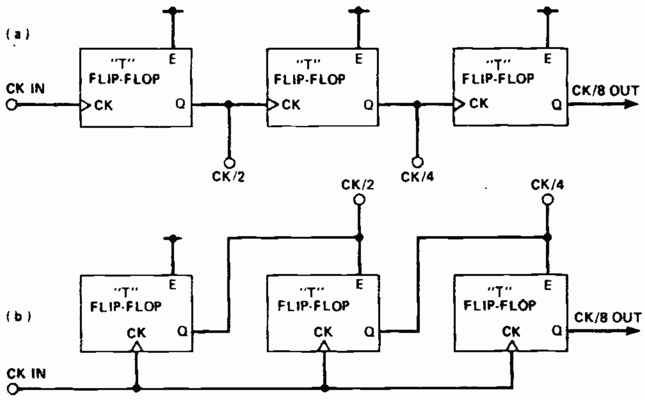


Fig. 3 (a) Asynchronous counter. (b) Synchronous counter.

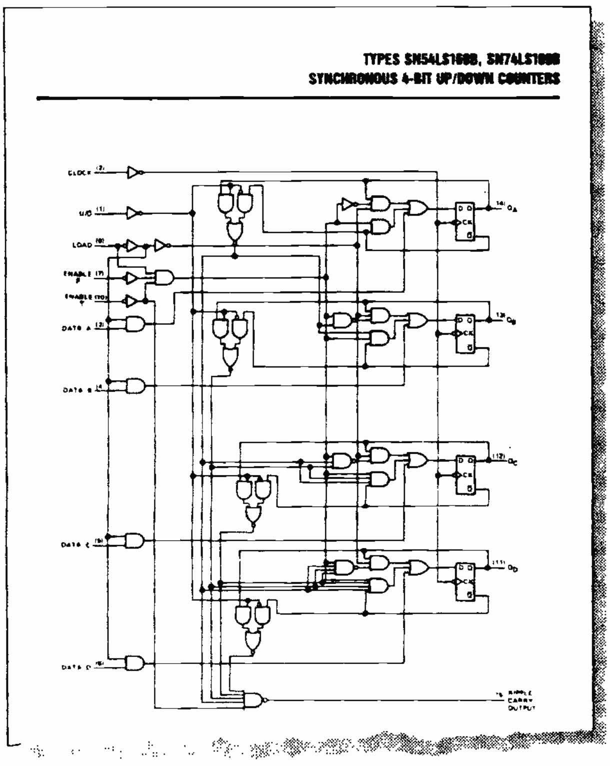


Fig. 4 The 74LS169B synchronous 4 bit up/down counter (courtesy of Texas Instruments).

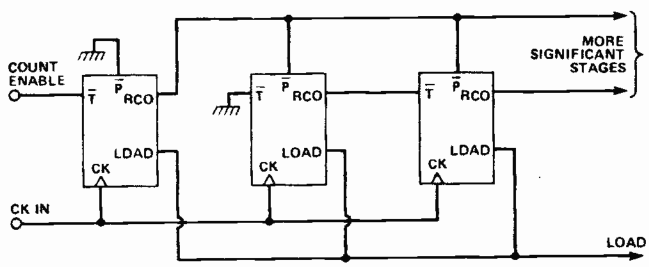


Fig. 5 Look ahead carry for synchronous counter cascading.

Within TTL, there are asynchronous (Fig. 3a) and synchronous (Fig. 3b) counters. Both use chains of bistable flip-flops, each of which divides by two the output from the previous flip-flop.

Asynchronous, or 'ripple' counters use the change in a preceding output to clock the following input, so that the propagation delays (the time between applying an input and getting a change at the output) of all the flip-flops are added together. This causes a phase shift, which depends on the number of cascaded flip-flops, and allows invalid outputs during the transition time between one count and the next. The synchronous counter, on the other hand, uses clocked flip-flops which are enabled or disabled by the outputs of preceding flip-flops, so that when a clock occurs, all enabled flip-flops change state simultaneously, and all disabled flip-flops do nothing.

From this it is obvious that for high modulo counters, synchronous operation is preferable. Most TTL counters are four bit (count of 16) building blocks, so we must physically cascade four bit packages to the modulo we require (about 1000), and care must be taken when doing this to ensure that the inter-package connections maintain synchronisation.

A good choice of counter device for this project is the 74LS169B from Texas Instruments (Fig. 4). This is a four bit binary up/down counter with a maximum speed of 35MHz (typical) and the important facility of 'synchronous loadability'. This means at any time a four bit data word can be presented to the chip, and it will be ignored until the next clock pulse after the LOAD pin is activated. Then the applied word will replace the previous state of the counter. This is crucial, as we will see later. It is worth mentioning in passing that the decade version of this counter (74LS168) is not a good choice, as it has invalid states which cause it to go out of sequence (you can load it with decimal 10 through 15, causing it to revert to its legal 0 through 9 via an unexpected route). However, when the counter is used as a free running maximal count divider, the decade counter has its uses.

The 74LS169B has a ripple carry output (RCO). This is effectively a decode of the terminal count of the device prior to the flip-flop inputs. Thus, as soon as the next state of all counter outputs is low (count zero) when counting down, or as soon as the next state of all outputs is high (count 15) when counting up, the ripple carry output is low. This condition persists until *after* the next clock. At all other times the RCO is high. Used with this output for cascading are two active-low enable inputs, P and T. P enables the counter chain, and T enables both the counter chain and the ripple carry output.

These signals are very important for synchronous cascasding. If you simply take the ripple carry out of the first stage and connect it to the clock input of the next, you create an asynchronous stage in your counter. The proper connection for cascading is shown in Fig. 5. The ripple carry output of the first (fastest) stage is used to enable (P) all subsequent counters, and the ripple carry output of the second stage is connected to the T enable of the third stage, whose RCO is connected to the T of the fourth, and so on. A common clock drives all stages synchronously. The 1000:1 range we require is provided by three cascaded four-bit counters (in fact 4096:1).

Ignoring for the moment the LOAD facility, we have now created a high modulo free-running counter which will count up or down according to the state of its direction select pin. Each output of the counter provides a square wave signal which is half the frequency of the previous output, and the RCO of the last (slowest) stage provides a negative going pulse (width = input clock period) every time the whole counter chain reaches its terminal count (once in 4096 input clock cycles).

Programming The Counter

If we were to load the counter with a defined value, it would start counting from this point, so the number of input clock cycles between counter start (after the LOAD signal is released) and the output of the RCO from the slowest stage would be determined by the value we loaded. This shows us a way to program the counter. If we feed the slowest stage RCO back into the counter by connecting it to the LOAD pins of all stages (Fig. 6a), every time the whole counter rolls over, it will reload itself. The usable output from the divider is a short 'tick' from the RCO/load feedback loop every time the counter reloads. It is these ticks that are used as clocks or triggers to other systems.

It is simplest to operate in this programmable mode by counting DOWN. You then have to load the value you want (instead of the difference between this and the max. count).

This design is really simple, and looks very elegant. There is, however, one tiny problem. If the slowest (most significant) stage of the chain is programmed to zero (when using the faster ranges of the divider) the RCO of this stage goes permanently low, inhibiting the whole chain by enforcing a constant LOAD condition. This imposes a severe limitation on the range of the pulse generator, by completely blocking all but the slowest ranges. The solution is shown in Fig. 6b. The RCO outputs of all stages are combined (logical AND) so only when all states are rolling over is a LOAD pulse generated.

As all signals are active low, an OR gate performs the necessary function by DeMorgan equivalence. OR gates with more than two inputs are not readily available in TTL, so the neat answer is to use a three input NOR gate (74LS27) followed by an inverter (74LS04). Note, though, that the propagation delays of these devices in the feedback loop will limit the top end performance of the generator, so we will in due course consider the use of faster devices for this design.

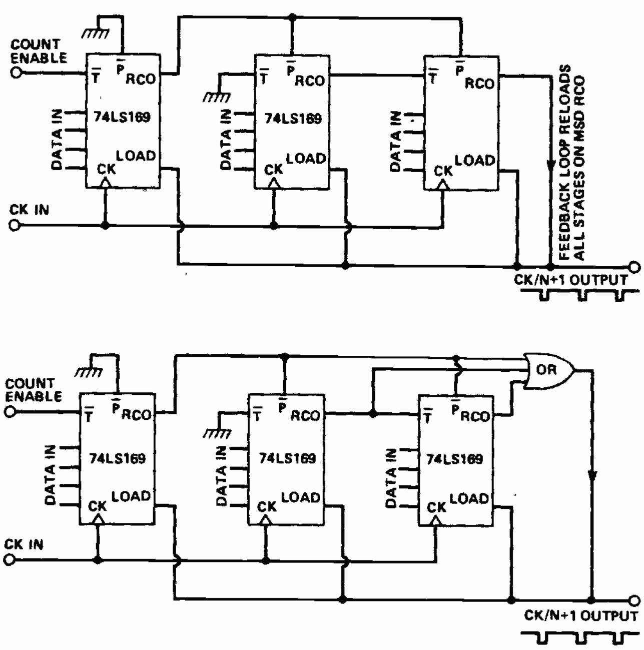


Fig. 6 (a) First attempt at programmable divider. (b) The improved circuit.

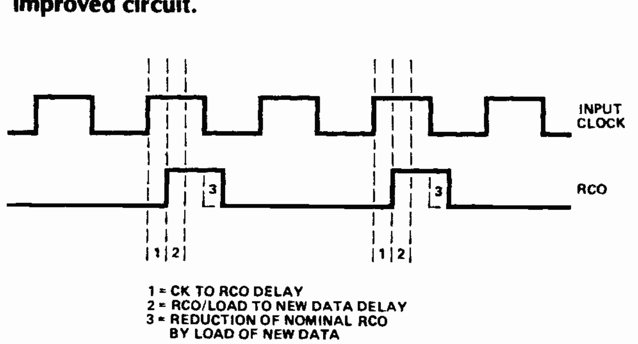


Fig. 7 Propagation delays in the divider chain.

Synchronicity

In these programmable divider chains, synchronous loading is all-important. If the LOAD was asynchronous (happening immediately the RCO signal appeared at the LOAD pins, as in the 74LS191) the counter reload would happen early (between two clock pulses), and the precision of the programmable divider would be spoilt.

As it is, the request to reload is recognised synchronously by the counter at the next clock pulse following the appearance of the RCO signal. This clock pulse, by loading the counter with the user's value, cancels the terminal count causing the RCO signal to go away. The propagation delays in the devices allow the RCO signal to take effect before it is cancelled (Fig. 7). As the clock following the terminal count reloads the counter and counting resumes on the clock following that, the counter modulo is always effectively one more than the loaded user value.

This is no problem when a microprocessor is used to load the modulo, as software is used to subtract one from the intended value before loading it into the counter. However, if we were to use switches, we would have to remember to subtract one 'by hand'.

The important outcome of this is the shortest delay sequence of the counter chain is divide by two. This is generated by loading the least significant counter stage with one and all more significant stages with zero. Loading all stages with zero is still a special case where all counting is inhibited, and this can be used as a stop condition for our pulse generator.

A single programmable counter chain provides us with a variable frequency square wave oscillator (Fig. 8). The RCO output of the final counter stage is fed back to the LOAD pins (via the NOR and invert gates), and also used to trigger a toggle flip-flop. The output from this is a good square wave with a frequency half the repetition rate of the RCO Pulses. This means that for a maximum output frequency of 5MHz, a 10MHz clock oscillator must be used.

Note that some alternative sources of 74LS169 may not operate in this closed loop (with feedback) configuration with a 20MHz input clock. 10MHz is safe, giving a 2.5MHz maximum frequency, or 400nS resolution, but it will prove necessary to go to 74AS or 74F series devices for 10MHz output. Special layout precautions are needed for these families of devices, and I will discuss them in due course. For the time being, the prototype may be run from, say, a 4MHz clock, at which the divider is absolutely guaranteed to operate with a 1μs resolution. For simplicity I have

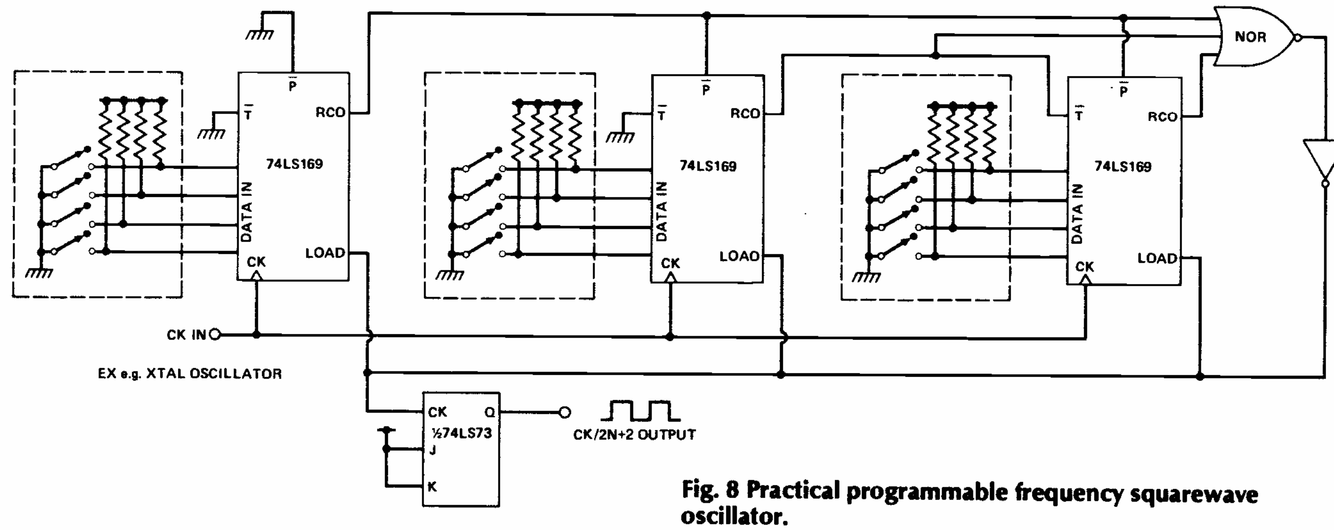


Fig. 8 Practical programmable frequency squarewave oscillator.

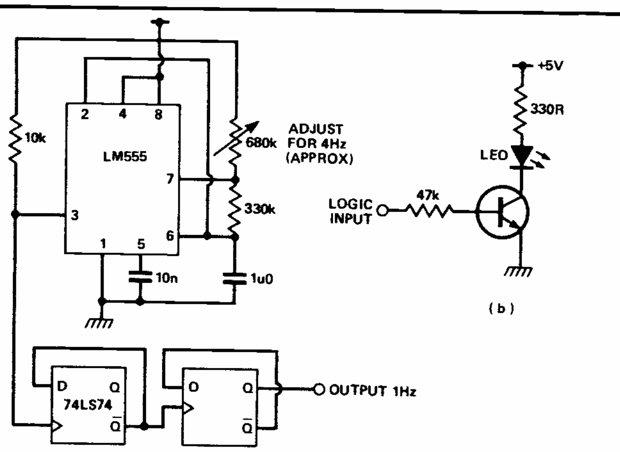


Fig. 9 (a) Slow oscillator for visual speed logic sequencing. (b) LED logic monitor.

shown counter programming by switches, but of course in the final implementation, the holding registers will be loaded by a microprocessor.

This is probably a good point to pause and say 'get experimenting'. Next month I will deal with the mechanisms of the pulse delay and width controls for the generators, but in the meantime, try wiring these circuits up on a breadboard. The only test gear you will need is a moderate 'scope (about 10MHz would do for the 4MHz version), or alternatively, you could rig a very slow oscillator using something like a 555 timer (Fig. 9a) and run the prototype at 1Hz. A logic probe or an LED with driver (Fig. 9b) can then be used to monitor the system.

ETI

CELLULAR RADIO

ETI's communications correspondent Keith Brindley casts an eye over cellular communications systems in the UK.

Whatever your point of view, modern electronic communications systems are pretty clever. How we use (or abuse) them is a different matter entirely. Personally, I can imagine little worse than the 'advantage' of a 'phone in my car. I have problems enough at home with the telephone ringing constantly, without having to suffer the same when I drive. Furthermore, I'm committed sufficiently as a serious motorist to feel the distractions of using a 'phone (or any means of telecommunications) in transit are dangerous and should not legally be allowed. Nevertheless, given the fact many people require telecommunications in order to do business, there is a need for a system such as cellular radio.

Mobile communications is not new. Mobile radio has been used by the military for nearly eighty years, and commercial organisations have been in operation providing systems to emergency services such as fire, police and ambulance for nearly as long. There are four other main types of mobile communications, which we should briefly consider before looking in depth at cellular radio itself.

Radiopaging. A call (generally digitally coded) is transmitted on a single frequency radio channel to a large number of receivers (sometimes nicknamed 'bleepers'). The call is headed by a receiver-specific code, triggering only one of the bleepers which correspondingly bleeps or otherwise indicates that a call has been received (Fig. 1). The call originator uses a normal telephone to dial the number dedicated to the receiver much as if the receiver is a telephone itself. However, the processing exchange routes the call to the receiver via a radio transmitter. In some sophisticated radiopaging systems a complete message is transmitted to the receiver and the user reads this on an LCD or similar display. Bleepers of more simple systems merely inform the user that a message is waiting. The user must then call a dedicated number via an ordinary telephone to receive the message.

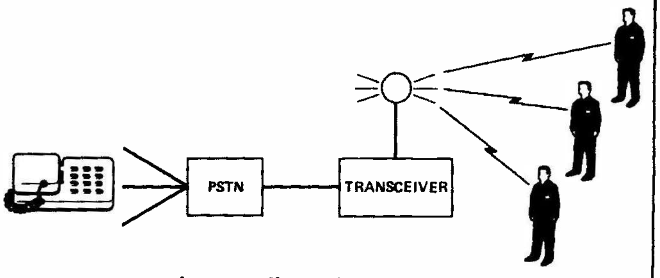


Fig. 1 Radiopaging.

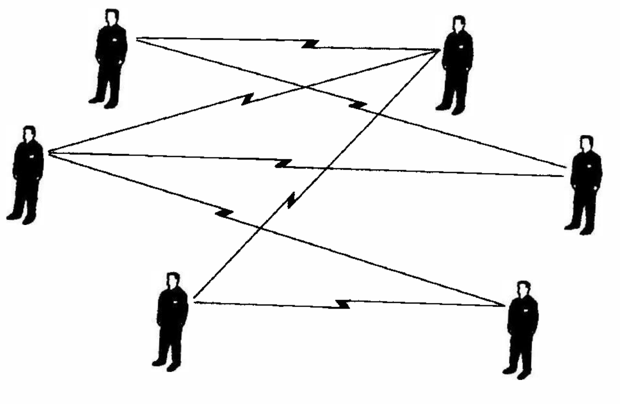


Fig. 2 Radiophone and CB radio.

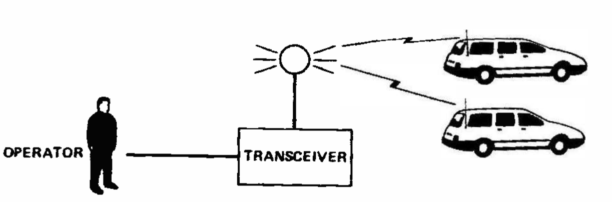


Fig. 3 Dispatch radio services.

Radiophone. Operating along the citizens' band lines, radiophones or 'walkie-talkies' as they are sometimes nicknamed are simple mobile receiver-transmitters, either hand-held or fixed in a car or base station. These use a single transmission frequency for both reception and transmission (Fig. 2). Use of a single frequency means only one user can transmit at a time, the other user in the communication link must wait until the first has finished before a reply can be made. This one-at-a-time operation is known as 'simplex' communication.

The bands of frequencies which the services use define the type of radiophone system. Citizens' band, for example, is so called because the service uses a band of frequencies which has been allocated for public citizens' use.

As a service, radiophone systems have significant drawbacks in addition to their simplex mode of operation. Users often have to queue, waiting for a free channel on which to transmit. This happens, particularly in urban areas where large numbers of users reside, or at busy times. Privacy is also impossible. Anyone with similar equipment can listen into a communication link.

Dispatch radio service. This is the sort of communications system used in taxis, ambulances, service vehicles and so on, in which a central or base station transmits and receives messages to and from mobile users (Fig. 3). The mobile users generally communicate only with the operator at the central station although some systems allow calls to be patched through from one mobile to another by the operator.

The disadvantages are fairly obvious. The number of mobile users on any one system is limited to the number of calls the operator can handle. Communication is simplex. Mobile users' transmitters can only be low powered so the distance over which the system

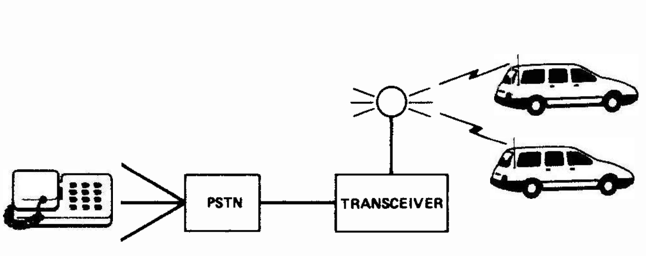


Fig. 4 Radio telephone services.

can be used is limited. Finally, with this service privacy is impossible.

Mobile telephone service. Commonly called radiotelephone services, the mobile telephone is linked over the air to the public switched telephone network (PSTN) so communication links with ordinary telephones can be set-up (Fig. 4). Communications are 'duplex', each user in the link able to talk at the same time.

This would seem the ideal mobile communications system. However, use of the available transmission frequency bands (HF and VHF) is not particularly economical and often a system may be busy, so long waits for a free channel can be expected.

A significant problem arises in the distance which a mobile user is from the central receiving station, and the consequent transmitter power. The further away the user is, the more powerful the transmitters must be. However, the more powerful the transmitters are the wider apart they must be to prevent interference with other transmitters of the same frequency or the fewer the available transmission frequencies. On the other hand, if low powered transmitters are used there is a good chance the mobile user will stray out of the transmission area during the call (the user *is* mobile, after all). Typically, quite powerful transmitters are used, with a limit to the number of channels available.

Privacy is also not possible in a standard system, although specialised systems can incorporate speech encryption — at a high cost.

One Better

It was the general poor quality of mobile communications and radio channel congestion which, in 1982, prompted the Government to allocate 30 MHz of radio frequency spectrum to a new mobile radio system, known as 'cellular' communciations. The system chosen was the 'Total access communications system' (TACS), a derivative of the already existing 'Advanced mobile phone service' (AMPS), developed by AT&T in the US.

Cellular communications systems are essentially an extension of existing mobile services. However, the use of computer controlled technology allows far greater economy of transmitting frequency bands and far greater efficiency. As far as the mobile user is concerned the receiving/transmitting apparatus is effectively an ordinary telephone, with all the usual telephone features (and more besides) and as easy to use.

Computer control of the switching apparatus means tremendous economy of the available frequency bands can be obtained. In turn, this means many more communication links can be made in the same number of frequency bands when compared with other mobile communications systems.

Finally, instead of a small number of relatively high powered HF or VHF transmitters being used to cover an area (as in radiotelephone systems), cellular communications systems use a very large number of low powered UHF transmitters. Again, it's computer control which allows this to work — calls can be automatically switched from one transmitter to the next as the mobile user moves between different transmitter reception areas.

We will look at each of these features of cellular communications in more depth.

Transmitter Power

Transmitter coverage in a mobile telephone service may be a circle with a radius of x km (Fig. 5). Outside this coverage zone is an interference zone, with a radius of about 5x km, in which the same radio frequencies used in the coverage zone cannot be used by another transmitter, otherwise interference of the two transmissions will probably occur. The same frequencies can be used outside the interference zone without interference occurring.

The key to the use of cellular communications is in the reduction in size of the coverage zones of the base stations (the receiver/transmitter) used. A smaller coverage zone means that fewer users will require the use of the base station. As a direct result of this, fewer channels are required to serve those fewer users.

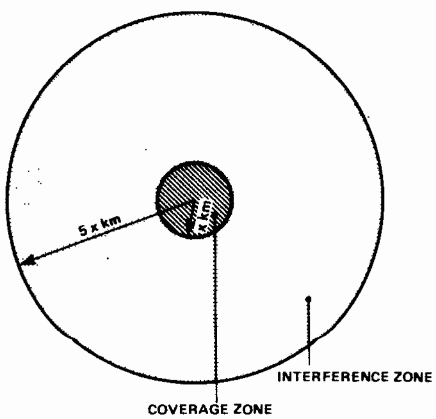


Fig. 5 Transmitter coverage.

In a cellular communications system the coverage areas are so small that they can be looked upon as 'cells' (hence the name) arranged in groups (known as 'clusters') over the land. Figure 6 shows seven-cell and 12-cell cluster arrangements. Others, such as four-cell and 21-cell arrangements, are also possible. Although each cell has a roughly circular coverage zone, they are usually thought of as hexagonal in shape and are much easier to draw because of this. Numbers within the cells merely indicate a particular frequency set of channels which transmitters in that cell use. Clusters are identical, in that correspondingly numbered cells in adjacent clusters use the same transmission frequencies.

The seven-cell cluster arrangement is ideal in terms of interference zones, as each base station is approximately five times the radius of its coverage zone away from the nearest base station with the same transmission frequency (measure it if you don't believe it!). So interference is unlikely. Any fewer cells in a cluster and corresponding base stations may be too close (and interference might occur). On the other hand, any more cells in a cluster mean that corresponding base stations would be further away than necessary and so transmission frequencies (and hence channels) are apparently being wasted.

However, the number of cells used in a cellular communications system depends on future requirements of the system, too. If more users than an existing arrangement can cope with wish to use the system, then two options are available.

Firstly, each cell can be split into seven smaller cells (ie, the cell itself becomes a cluster) so base station coverage zones become smaller and a greater number of users can be accommodated.

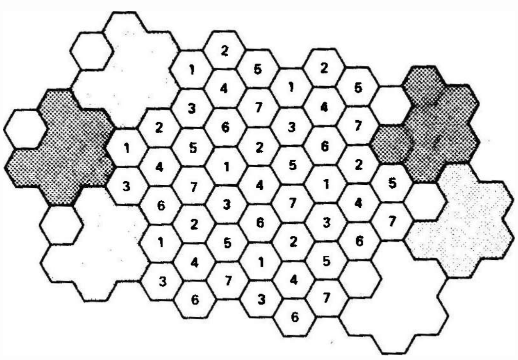
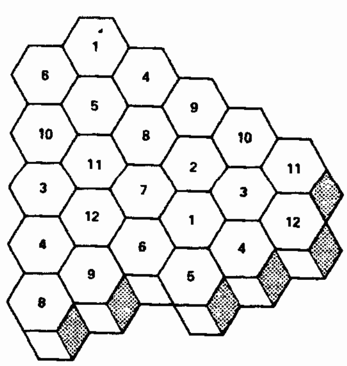


Fig. 6 (a) 7 cell repeating pattern.



(b) 12 cell repeating pattern.

The second alternative is for each cell to be sectored by using directional aerials which only cover one part of the cell. For example, three aerials with 120° each will cover one cell, using the same transmission frequencies, effectively tripling the number of channels which can be used and giving a corresponding increase in the number of possible users.

The two cellular communications systems in use in the UK, Cellnet and Vodafone, illustrate this well. Vodafone uses the seven-cell arrangement, but in urban areas has sectored each cell with three 120° directional aerials, which is about the limit of the system. Cellnet, on the other hand, has a 12-cell arrangement and so can sector its cells more thoroughly (with six 60° directional aerials in each cell) before reaching the system limits. After the sectoring limits are reached, further users can only be incorporated by splitting each cell into seven or 12 smaller cells.

Future Limits

However, there is a lower limit to cell size when splitting, which depends primarily on system accuracy and where the aerial can be sited. The smallest usable cell (in London, actually) is about 2 km in radius. Smaller than that and it becomes too critical to be able to site an aerial in a convenient position. It's impossible to site an aerial, say, smack in the middle of Tottenham Court Road if that's the ideal spot. Tolerances of a few hundred metres must be built-in to the system to allow convenient aerial siting, which inevitably would cause interference problems unless cell size is sufficiently great. So, in effect, there is a maximum limit to the number of users which can be accommodated on a system.

When the limit is reached will depend on the rate of growth of new users to a system. Eventually, however, the time must come when both existing cellular communications systems are choc-a-block and new systems will have to be built. Even then, unless far more frequency spectrum is allocated to cellular communications, it's not going to be the case that we'll all have our very own cellular telephone in the future — the wired 'phone will be with us for a long time yet!

And talking of the future, it is interesting to note that the UK TACS systems, Cellnet and Vodafone, are each licensed to use only 300 of the 1,000 channels allocated by the Government for present cellular communications purposes (21 of these are allocated for control data purposes, leaving 279 channels each). The remaining 400 are presently allocated for the Europe-wide second generation cellular communications system, planned for the 1990s.

The older mobile telephone services use HF and VHF transmissions which have unpredictable and long propagation ranges. Cellular communications systems, on the other hand, use UHF transmissions which are essentially line-of-sight (communications links can only take place up to a maximum of about 40 km and won't go any further than the horizon), and extremely directional, with highly predictable propagation ranges. Also, because UHF transmissions penetrate deeply into buildings, tunnels, and other physical barriers (unlike HF and VHF transmissions) lower power transmissions can be used, allowing ranges of only a few hundred metres or so. Radiotelephone services operating at HF and VHF must 'swamp' an area to provide the equivalent penetration, creating long propagation ranges.

For reference, the UK TACS cellular communications system uses transmisions at typical effective radiated power (ERP) levels of 50 to 1000 watts for base stations, and 0.6 watts to 10 watts for mobile equipment. The frequency band 935 to 960 MHz is used for 'forward direction' transmissions (from the base station to the mobile) and the band 890 to 915 MHz is used for the 'reverse direction' transmissions from the mobile to the base station).

Cell size is not fixed throughout the country, as you've probably gathered from this discussion already. In rural areas a cell can be quite large (up to 40 km in radius, as we've seen), while in urban areas cells are decidedly small. The general rule-of-thumb is that all cells should have about the same number of users and so have the same number of channels available for those users.

Over To You

The use of quite small cells in a cellular communications system leads to the problem which occurs when a mobile user leaves one cell and enters another. The problem is solved with the use of complex network switches which automatically switch the links through the best base station so that reception and transmission is maintained. The complete national system (Fig. 7) is based around these network switches, which route communications links in the first place and perform the functions of time-keeping and maintenance of tarrif data. Finally, the network switches monitor the position of all users so that any call to a mobile user will be quickly routed.

The process of switching from one base station to another when a mobile user goes from one cell to another (and indeed, from one sector of a cell to another) is known as 'hand-off'. This occurs in about 300ms so usually is undetected by the user. Individual cell base stations monitor each communications link in progress, and if communication quality falls below a predefined level the network switch is automatically alerted.

Base station monitoring is remarkably simple. Besides the channels available for telehone communication links, there are some channels reserved for control and signalling communications. The mobile equipment transmits control signals at regular intervals, but when actually in the process of a communications link, it also transmits a tone in a control channel. If the signal-to-noise ratio of the tone falls, indicating the user is moving out of the cell, the base station informs the nearest network switch. It is then up to the network switch to monitor signal strengths at all surrounding cell base stations and switch the link onto a different channel and base station.

Signal strength sensing of this type can also provide the mobile's position to the network switch, at least in terms of the base station which receives the strongest signal. This method of locating the mobile works best in the countryside where signal strength is usually dependent only on the mobile's distance from the base station. In an urban area signal strength sensing is not so useful. A mobile may produce a low signal strength even when comparatively close to a base station because it is being nulled by various fixed and mobile objects. Some cellular communications systems (not TACS) use an alternative form of positional sensing, called 'trilateration', in which the mobile regularly transmits pulses which are received by at least three base stations. The relative delays experienced by the base stations in receiving each pulse indicates the distance the mobile is away from each base station and so gives its position.

Network switches are connected directly to the PSTN so that any fixed telephone subscriber can call any mobile user, or vice versa, and any mobile user can call any other mobile user — world-wide.

Roaming

If a user moves throughout a cellular communications system roaming is said to occur. To roam right throughout the system means hand-off will at some time be required from one cell to another, where the cells belong to adjacent network switches. For this to happen, inter-switch hand-off is required, not just inter-cell hand-off as we've previously considered.

Interestingly, the Vodafone system has not yet got inter-switch hand-off (Cellnet has just been updated to incorporate it). However, Vodafone's network

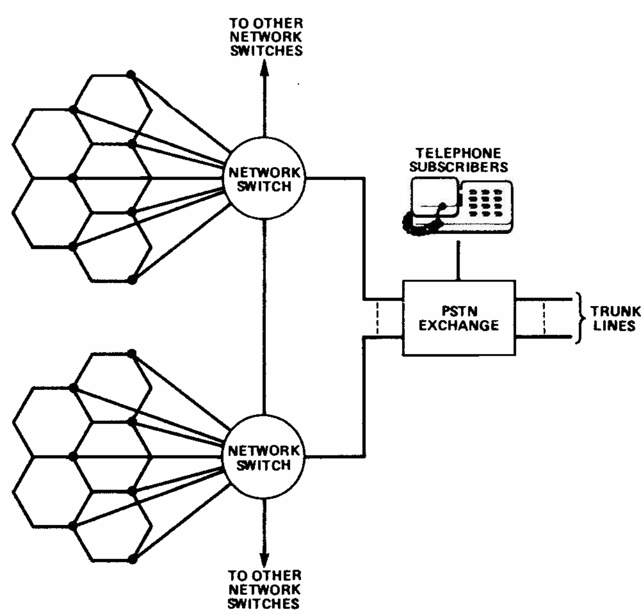


Fig. 7 Connection of cells to the public Switched Telephone Network through interconnected network switches.

switches cover very large areas (each switch can accommodate more than 100,000 users, compared with Cellnet's 30,000 users) so the facility is not so critical.

Features

Now we've described the system, what can it do? First and foremost cellular communications provides a means whereby a mobile user, in a vehicle or on foot, can maintain telephone communication with whoever the user wishes to contact, or whoever wishes to contact the user. So all the features usually taken for granted with the 'phone are available wherever the user travels.

More than this, there are some features which the existing landline telephone network cannot usually provide. These extra features include:

- **Call diversion.** The incoming call can be assigned to an alternative number when the mobile is busy, turned off, or does not reply within a certain period.
- **Three party conference.** Board meetings over the 'phone. A conversation between three or more parties can be heard simultaneously by all parties.
- **Call transfer.** A call can be transferred to another telephone location by the mobile user.
- **Selective call barring.** Certain types of call, such as international calls, can be denied to the mobile user.
- **Alarm call.** The network switch will automatically call a mobile at a time programmed by the user.
- **Camp-on.** If the mobile is in use, an incoming call will be put 'on hold' at the network switch, until the mobile is free. At the same time, the mobile user will be warned a call is being held.
- **Call holding.** An existing call can be put on hold, by the mobile user, while a camped-on call is answered.

Most mobile equipment also has the following facilities:

- **On-hook dialling.** The handset can be left on its cradle until the called party answers.

Cellnet only
British Telecom Mobile Phone Division
Docos House
68–82 Commercial Road
London E1
Tel: (01) 432-6280
Caledonian Communications Ltd
17 Waterloo Place
Edinburgh
Scotland
Tel: (031) 558 1144
The Carphone Company
Vallis House
57 Vallis Road
Frome
Somerset
Tel: (0373) 73220
Cellcom Ltd
Sutherland House
70-78 The Broadway
London NW9
Tel: (01) 202 2322
Cellular One Ltd
114 Windmill Road
Brentford
Middlesex
Tel: (01) 568 1166
Granada Mobile Telephones
Dane Street Chambers
Dane Street
Bedford
Tel: (0234) 55233
Marconi Mobile Radio
Marconi Communications Systems Ltd
New Street
Chelmsford
Essex
Tel: (0245) 353221
Motorola Communications Services Ltd
Jay's Close
Viables Industrial Estate
Basingstoke
Hants
Tel: (0256) 58211
Relcom Communications Ltd
Hammersmith Industrial Estate
Winslow Road
London W6
Tel: (01) 741 8581
Securicor Robophone Communications Ltd
Ambassador House
Brigstock Road
Thornton Heath
Surrey
Tel: 9)1) 689 2144
Tactico Ltd
65 Logie Green Road
Edinburgh
Scotland
Tel: (031) 556 6167/8

Racal Vodaphone only
Advance Car Telephones
226 Christchurch Avenue
Kilburn
London NW6
Tel: (01) 328 6514
Automobile Association
AA Vodaphone
Fanum House
52 London Road
Twickenham
Middlesex
Tel: (01) 891 6191
ECT Cellular Ltd
14 Muswell Hill Road
Highgate
London N6
Tel: (01) 368 4321
Ericsson Information Systems Ltd
18 Clerkenwell Road London EC1
Tel: (01) 251 0187
In-Car Telephones
Unit 19
Acton Business Park
The Vale
Acton
London W3
Tel: (01) 749 9572
London Car Telephones Ltd
9 Lower Addiscombe Road
Croydon
Surrey CR0 6PQ
Tel: (01) 680 4444
Racal-Vodac Ltd
Freepost
Newbury
Berkshire RG13 1BR
Tel: (0653) 69000
Storno Ltd
110 Grafton Road
London NW5
Tel: (01) 267 5136
Thorn-Ericsson Ltd.
Viking House
Boundary Lane
Horsham
West Sussex
Tel: (0403) 64166

Both Cellnet and Racal-Vodaphone
AirCall PLC
108 Rochester Row
London SW1P 1JP
Tel: (01) 843 9000
Answercall Cellular Telephones PLC
23 Mount Street
London W1
Tel: (01) 499 2476
Cellcall Ltd
5 St Botolph Street
London EC3
Tel: (01) 283 1122
Digital Mobile Communications Ltd
Station House
Harrow Road
Wembley
Middlesex
Tel: (01) 903 7753
Europa Communications Ltd
Merevale House
Parkshot
Richmond
Surrey TW9 2HW
Tel: (01) 940 9433
Excell Communications Ltd
Excell House
Hale
Altrincham
Cheshire
Tel: (061) 941 7006
Gazelle Group PLC
Ponton road
London SW8 5BL
Tel: (01) 627 0000
International Communications Ltd
Unit 19
Commercial Way
Park Royal
London NW10 1YP
Tel: (01) 964 0065

- **Short code dialling.** Frequently called numbers can be stored and recalled using a code (generally two digits).
- **Electronic lock.** Prevents unauthorised use.
- **Received call indicator.** The equipment indicates a call has been received while the user has been away from the vehicle.
- **Hands free operation.** Conversation can take place without lifting the handset.
- **Last number redial.** The previously dialled number can be redialled, by pushing a single button.

Quality

Overall quality is pretty good. However, it relies mainly on the aerial the user has on the car. It's no good using a bent piece of coat hanger wire and expecting first class results. In a high signal area, with the vehicle stationary you'll have a successful call, but start to move, especially through fringe reception areas, and quality will be lousy. There's no doubt, an incorrect, or even a correct but incorrectly fitted aerial will give poor results.

Research carried out by laboratories, cellular operators, and aerial manufacturers has shown the best position for an aerial is on the roof of a vehicle. So, don't expect to be able to use your existing car radio aerial for cellular purposes. If you don't fancy the idea of defacing your brand new Ford Chevalier RS Turbo GX with an ugly brute of a roof aerial, then cellular communication is not for you.

There are alternatives such as an 'on-glass' aerial or an elevated feed aerial. Unless these are properly installed, you may find you've wasted an awful lot of money. Consult the professionals. That does not necessarily mean the people you obtain the mobile equipment from. Although they will usually only sell you the correct equipment for your needs, they may not be as technically aware of all the aerial considerations as they could be.

The Bottom Line

Finally, how much does all this cost? You can't get something for nothing in this world, and what you get out of your cellular communications telephone will depend on how much you put into it. The 'phones themselves will cost you anything from about £1,000 to £2,500 to buy and from £35 to £70 per month to lease, depending on facilities.

Running costs depend on which of the two systems and which features you use. Typical costs (after the initial connection) run at around £25 per month for rental of service, and around 33p per minute of daytime calls.

From these figures, you'll appreciate that cellular communications is presently a company luxury or a rich man's status symbol. If you're stinking rich or a mobile executive and you think you need easier communications to do your job better, cellular communications could be for you. As time goes on and costs come down (which they are bound to do as the market grows) ordinary consumers may be tempted. For us ordinary mortals my advice is to wait for a year and see what happens. Already, prices are falling.

Some of the data used in this article was taken from Cellular Communications 1986 — A Worldwide Report, produced by Mintel Productions Ltd, and is reproduced courtesy of Cellnet.

THE MIDI INTERFACE

Alan Robinson continues his voyage of discovery in the turbulent sea of the Musical Instrument Digital Interface (MIDI). This month he looks at the confusing array of software codes used by the standard.

Last month I gave a general introduction to the MIDI Interface and described its hardware. In this article I shall investigate the software side of things — the actual meaning of the bytes sent from the transmitter to the receiver. Before diving in to the details, it must be stressed that not all equipment is capable of transmitting or interpreting, all the information the standard covers. Two instruments can only communicate at their lowest common level.

Fortunately, common ground usually covers the basics to the extent that a make A keyboard can drive a make B synth, even if some of its more subtle features do not work properly.

Status And Data Bytes

The bytes transmitted are divided up into two types, called status bytes and data bytes. Status bytes indicate the type of data being transmitted, whereas data bytes contain the data itself. Status bytes are distinguished by having bit 7 set. Data bytes have bit 7 reset. Some of the data uses more than seven bits, so is transmitted as a pair of bytes (low byte first), giving a total of fourteen bits of data.

Information transmitted is sent in batches of bytes called messages. Each message starts with a status byte (with bit 7 set) indicating the type of data about to be transmitted, and then a number of data bytes containing the data itself. However, there is one complication. To speed up transmission there is a feature called 'running status'. Basically, this means a status byte only need be transmitted when it differs from the one at the start of the last message. If a receiver receives data bytes which are not preceded by a status byte, it should assume that the status is the same as last time and the data is more of the same. For example, the transmitter might send information requiring that a chord of three notes be played. This is done by transmitting three 'note-on' messages, but the note-on status byte may be omitted for the second and third messages.

Channel Messages

There are two basic types of message: 'channel' messages, and 'system' messages. Channel messages relate to instrument sounds or voices. They are called channel messages because a transmitter can send out information which is destined for more than one receiver. A sequencer, for example, might control more than one instrument. Each receiver is assigned to a channel number and only responds to messages addressed to it.

MIDI caters for sixteen channels. As far as the user is concerned these are numbered from 1 to 16, but they are actually coded as binary values from 0 to 15. The binary number transmitted is always one less than the user's idea of the channel number — another trap for the unwary. Assignment of channel numbers is covered later. Suffice it to say for now that the channel number is contained in the least significant four bits of the status byte at the start of a channel message. Receivers must examine these four bits to decide whether or not a message is addressed to them, and ignore it if not. Channel messages are further divided into channel 'voice' messages and channel 'mode' messages.

Channel Voice Messages

Channel voice messages are the most frequent. These allow one piece of MIDI eqipment (typically a keyboard or sequencer) to make another piece of equipment (typically a synth) to play something. This includes turning notes on and off, but also gives control of pitch bend, and modulation and many other aspects of the sounds produced as well.

Table 1 summarises the channel voice messages. The last four bits of the status byte form the channel number (shown as **n**) and the most significant bit is always 1 to distinguish it from data bytes. The actual function of the status byte is coded into bits 4, 5, and 6. Of the eight possible values of these three bits, seven are used for channel messages: **8n**, **9n**, **An**, **Bn**, **Cn**, **Dn**, and **En**.

Table 1 Channel voice messages

Status Byte Binary	Hex	Description	Date Bytes
1000nnnn	8n	Note off	1. key number 2. release velocity
1001nnnn	9n	Note on	1. key number 2. key velocity
1010nnnn	An	Polyphonic key pressure	1. key number 2. key pressure
1011nnnn	Bn	Control change (see tables 2–4)	1. control number 2. control value
1100nnnn	Cn	Program change	1. program number
1101nnnn	Dn	Channel pressure	1. channel pressure
1110nnnn	En	Pitch bender	1. pitch bend (LSB) 2. pitch bend (MSB)

Status messages **8n**, and **An** give information relating to individual keys on the keyboard. The first data byte transmitted in all three cases is the MIDI key number. The key numbers range from 0 to 127, so that in theory MIDI can handle keyboards of up to 128 notes, or a span of well over ten octaves.

Middle C is defined to be key number 60 (or 3C in hex), so that key number 0 is the C five octaves below middle C, which at concert pitch has a fundamental frequency of about 8.2Hz. Needless to say it is not used too often! Key number 127 is the G five and a bit octaves above middle C, with a frequency of 12.5kHz at concert pitch. That is not used often either!

Status message **9n** is the 'note-on' messsage, which is generated when a key in pressed. The first byte transmitted is the key number. The second byte is the key velocity, or how hard it was struck. Many cheaper keyboards are not velocity sensitive, and these transmit a medium value of 64 (40 hex). The exact meaning of the key velocity byte is not defined.

The opposite of the **9n** note-on message is the **8n** 'note-off' message, which is generated when a key is released. This is very similar to a 'note-on' message in that the first data byte transmitted is the key number. The second data byte is the release velocity. As with the key velocity in note-on messages, the release velocity in the note-off message is not precisely defined. Keyboards without velocity sensing again transmit the median value of 64 (40 hex). Every note-on message must be followed eventually by a corresponding note-off message.

There is another potential trap for the unwary with note-on and note-off messages. A note-on message with a key velocity of zero is defined to be a note-off event instead. In this case the receiver has to assume that the release velocity is median value of 64 (40 hex), since this cannot be given. This special case allows note-on and note-off information to be transmitted in the running status mode without re-issuing the status byte, which can sometimes speed things up a bit. Some keyboards will always transmit note-off messages this way though, so beware!

The third kind of individual key related message is the **An** 'polyphonic key pressure' or 'after touch' message. Again the first data byte sent is the key number. The second is the pressure value. The specification says nothing about the relationship between pressure and pressure value sent but in practice a value of 0 to 1 means virtually none, and a value of 127 means loads of it.

The polyphonic key pressure message is not often used because only the most expensive keyboards have pressure sensing for individual keys. A more common and closely related message is the **Dn** status 'channel pressure' message. This is the same except that the key number is not transmitted, so the message contains only one data byte. The data represents the force applied to all the keys.

Status message **En** is for the pitch bender. In this case the data transmitted is a 14-bit number, transmitted as two data bytes containing seven data bits each. The median value in this case is 8192, or 2000 in hexadecimal, which is transmitted as the three hexadecimal bytes, En 00 20. This value corresponds to a pitch bend of zero. Again the MIDI specification is not specific about the relationship between the pitch bend value transmitted and the amount of pitch bend it corresponds to. Often instruments only have seven bit resolution, with the least significant byte containing only dummy information, although fourteen bits of data are always transmitted.

The two different kinds of key pressure message and the pitch bender message are unlike the note-on and note-off messages in that they are continuously variable values which must be repeatedly transmitted many times. The MIDI specification makes no mention of the rate they are transmitted at. This is up to the equipment designer. In practice, these messages are only transmitted when the pressure or pitch bend value changes so transmission rate depends on the resolution of the sensors, the rate they are scanned, and their use at the time.

Status message **Cn** is the 'Program Change' or 'Patch Change' message. These messages contain a single data byte which selects one of 128 'programs' or 'patches', numbered from 0 to 127. Each program is a complete set of control values that define a particular instrument voice or sound. Program change commands instruct the receiver to switch to the new voice selected by the data byte, assuming of course this is something the receiver is capable of. Whenever a new program is selected on an instrument's controls, a program change message is transmitted.

The final channel voice message is the **Bn** 'control change' message used to transmit control parameter changes other than the key pressure and pitch bender values. Each control change message contains two

Number	Function
0	Continuous Controller 0 (MSB)
1	Continuous Controller 1 (MSB)
2–31	Continuous Controllers 3–31 (MSBs)
32	Continuous Controller 0 (LSB)
33	Continuous Controller 1 (LSB)
34–63	Continuos Controllers 3–31 (LSBs)
64–95	On/Off Switches (0 = off, 127 = on)
96–121	Undefined
122–127	Reserved for Channel Mode message

Table 2 Assignment of control numbers

data bytes, the first giving the control number and the second the control value. Table 2 summarises the uses of the control numbers according to the MIDI specification.

Controller numbers 0 to 63 are assigned to 32 'continuous' controllers such as potentiometer inputs. These controllers can have 14-bit resolution if required. One controller number in the range 0–31 is assigned to the most significant byte (MSB), and another in the range 3–63 is assigned to the least significant byte (LSB). The LSB for a controller may or may not be transmitted or received, depending on whether or not seven bit resolution is enough for its function, and entirely at the discretion of the manufacturer.

Controller numbers 64 to 95 are assigned to switches, which can be either on or off. To turn a switch off, a value of 0 is sent, and to turn it on, a value of 127 is sent. How a receiver interprets any other value (1–126) in between is not defined.

Controller 1 is assigned to the modulation wheel. For the others the MIDI specification simply says they may be assigned by manufacturers as they please. An earlier, now out of date, MIDI specification assigned controller 0 to the pitch bender, but this is now done by **En** status messages.

This vagueness about the assignment of the controllers is unfortunate, but something of an unofficial standard has emerged. Table 3 summarises these commonly used assignments.

Number	Function
1	Modulation Wheel
2	Breath Controller
4	Foot Controller
5	Portamento Time
6	Data Entry Knob
7	Main Volume
64	Sustain Pedal
65	Portamento
66	Sostenuto
67	Soft Pedal
96	Data Entry Increment
97	Data Entry Decrement

Table 3 Common use of controllers

Message	Status	Data
Local control off	Bn	122,0
Local control on	Bn	122,127
All notes off	Bn	123,0
Omni Mode off	Bn	124,0
Omni Mode on	Bn	125,0
Mono Mode on	Bn	126, No. of channels
Poly Mode on	Bn	127,0

Table 4 Channel mode messages

Channel Mode Messages

Control numbers 122–127 are assigned to an entirely different class of message: Channel 'mode' messsages. These are used to define the operating mode of the voices for a particular setup. They are summarised in Table 4. Control number 122 can be used to disconnect an instrument's keyboard from its sound generating circuits so that the keyboard data goes only to the MIDI OUT socket, and the sound generators are only controlled by the MIDI IN data received. This is an optional feature. Not all instruments are capable of being controlled in this way.

Control messages 123 to 127 are all 'All notes off' commands, 123 being different in that it serves no other purpose. However, the MIDI specification says that in no case should these be used to turn off notes previously turned on by 'note on' messages. These should be turned off by specific 'note off' commands. Since there is no possibility of notes otherwise being left on, 'All notes off' commands may safely be ignored. Readers puzzled by this strange aspect of the specification may rest assured they are not alone!

Each instrument is assigned by the user to a channel number, called the instrument's 'Basic Channel', using its own controls. Control messages 124 to 127 control the operating mode of the instrument's voices. These messages control two switches, the OMNI ON/OFF switch and the POLY/MONO select switch.

OMNI ON, POLY mode. In this case voice messages are received on all channels and routed to the voices according to the instrument's own internal polyphonic assignment algorithm. Voice messages are only transmitted on the basic channel.

OMNI ON, MONO mode. In this case voice messages are received on all channels but only routed to one voice, monophonically. Voice messages for one voice only are transmitted on the basic channel.

OMNI OFF, POLY mode. In this case voice messages are only received and transmitted on the basic channel, and received messages are assigned to voices according to the instrument's own polyphonic algorithm.

OMNI OFF, MONO mode. In this case voice messages are received on a number of channels, and each receive channel is assigned to one voice. The number of channels it receives on is given in the last data byte of the 'mono select' message it receives. If, for example, an instrument assigned to basic channel 5 received a MONO mode message specifying four channels, it would receive voice messages on channels 5, 6, 7, and 8 (four channels starting from its basic channel). Each of the four channels would be assigned to one of its voices, assuming of course it has four voices to assign. A special case is where the MONO mode message specifies zero as the number of channels to receive on, when it will assign itself to as many channels as it has voices, again starting from its basic channel number. An instrument in this mode combination will transmit voice messages for a number of voices on the same channel

Not all instruments are capable of operating in all these modes. Some are assigned to fixed basic channel numbers the user has no control over. On power up, all instruments should default to OMNI on, POLY mode. It is also normal for an instrument to initially assign itself to basic channel 1 until told otherwise by its operator.

System Messages

System messages are unlike channel voice and mode messages. They are not related to channels and do not have channel numbers included in the status byte. Particularly important, the system messages include the codes that are used to synchronise sequencer timing so a synth and a drum machine can play along together without any voice messages being transmitted between them. Each sequencer is programmed separately and only the timing codes are needed to make them play together.

System messages are distinguished by status bytes with the four most significant bits set (**Fn**). They are divided into three types, 'System Common,' 'System Real Time', and 'System Exclusive'. Table 5 summarises the system messages.

System exclusive messages are meant for a particular manufacturer's equipment and can contain any data the manufacturer chooses to put in. The status byte is **F0**, followed by a number of data bytes, the first of which is the manufacturer's ID number. ID numbers are allocated for registered manufacturers by the MIDI Association (for a price!).

The message is terminated by an **F7** 'end of exclusive' status byte, or any other status except real time codes **F8** to **FE**. An instrument which does not recognise the ID code at the start of a data stream ignores the data until another status byte indicates the end of the exclusive data.

Table 5 System messages

	Status	Data	Function
Exclusive	F0	ID data data	Data specific to a particular manufacturer's equipment (any number of data bytes)
Common	F1	–	Undefined
	F2	LSB, MSB	Song Position Pointer
	F3	Song No.	Song Select
	F4	–	Undefined
	F5	–	Undefined
	F6	none	Tune Request
	F7	none	End of Exclusive
Real time	F8	–	Timing Clock
	F9	–	Undefined
	FA	–	Start
	FB	–	Continue
	FC	–	Stop
	FD	–	Undefined
	FE	–	Active Sensing
	FF	–	System Reset

There are four system common messages. **F3** status messages are 'song select' messages, which choose between songs, numbered 0 to 127, and **F6** 'tune request' is used to start analogue synthesiser automatic oscillator tuning. **F2** status messages are 'Song Position Pointer' messages. These messages are closely related to the real time 'timing clock', 'start', 'continue', and 'stop' codes, and allow the user to start or continue a sequence from a specified point in the song.

The data given is a fourteen bit value coded as two bytes, the least significant being sent first. It indicates a time within the song in multiples of sixteenth notes, or semiquavers, from the beginning of the song. Very few instruments actually implement the song position pointer feature.

There are six system 'real time' status codes. These do not have associated data bytes, and with the exception of system reset code **FE** can be transmitted at any time, even in the middle of other messages. Another little pitfall for the unwary! This aspect of the specification is to allow greater timing accuracy in their transmissions, though this is only really important in the case of the timing clock **F8**. 'Running status' is not changed by the system timing codes.

F8 timing clocks are sent at the rate of 24 clocks per quarter note or crochet, and are used to synchronise sequencers and instruments with built-in sequencers, a synthesiser sequencer and a drum machine, for example. Timing clocks must be ignored until an **FA** 'start' or **FB** 'continue' code is received. This is important to note, because some instruments continuously transmit timing clocks regardless of whether their sequencers are running or not. **FC** 'stop' codes stop the sequence running.

The 'start' code starts the sequence from the beginning every time, whereas the 'continue' code starts it from where it was last stopped, or if the 'song position pointer' feature described above is used, from where the pointer was set to.

FE codes are for 'active sensing'. This means some transmitters continuously produce this code when nothing else is being transmitted, just so the receiver knows the transmitter is still active and plugged in. The transmitter should transmit this code at least every 300ms (and preferably faster, say every 100ms to 250ms) when there is no other activity on its output. A receiver which does not receive any input for at least 300ms (or perhaps 400ms to be on the safe side) then knows that the transmitter is not longer active and can turn off its voices and return automatically to stand-alone operation. However, not all transmitters produce this code so receivers should not assume its presence until an **FE** code is actually received in the first place.

The remaining real time code is 'system reset'. This is used to reset the system to the condition of just having been switched on. It should be used sparingly, if ever at all. Instruments must not send out this as part of their cold start routines because this could result in two connected back to back forever resetting each other when switched on and never actually getting to the stage where they could be used!

In some cases it does not seem to make much sense for an instrument to completely reset itself just because a MIDI command has told it to, especially in the case of something like a computer used as a sequencer. Does the user really want it to completely re-boot the sequencer software and quite possibly lose its programmed sequence because its MIDI IN socket told it to? Perhaps the sequencer software disk is no longer in the drive and the computer can then no longer even be a sequencer without changing disks. Not very helpful! In this case a decidedly warm start would make more sense than a cold start. This is one aspect of the MIDI specification that should perhaps be given a somewhat loose interpretation!

ETI

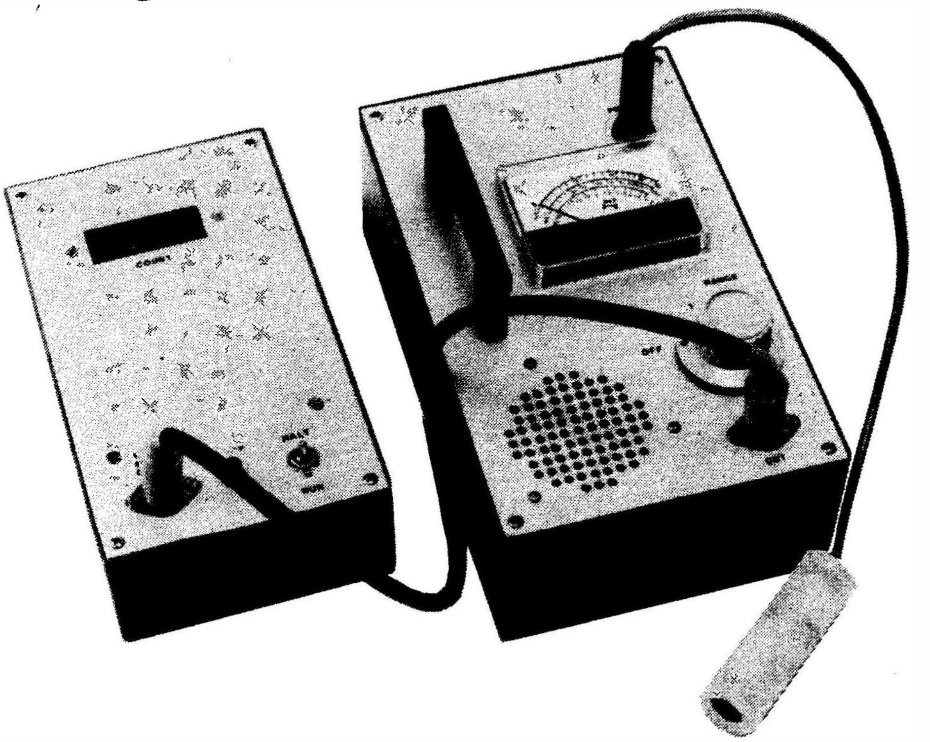
GEIGER RATEMETER AND COUNTER

Colin Seymour describes a cheap, practical radiation monitor which can be used both for instant dose-rate checks and for long-term, accurate counting.

With the present concern about nuclear radiation, especially since the Chernobyl disaster and the sinking of a nuclear submarine, it is not surprising that many people have been trying to obtain Geiger counters or other radiation monitoring devices. Sales of portable detectors have increased and ETI and other magazines have received countless requests for information on radiation measurement and reprints of past articles describing detecting equipment.

We published a basic circuit for a Geiger ratemeter soon after the Chernobyl accident (see p.17, ETI July 1986). Since then several other electronics journals have published ratemeter designs of various levels of sophistication.

The design presented here is the result of a considerable period of development and offers excellent performance and versatility at low cost. It consists of two units, a self-contained Geiger ratemeter and a separate pulse counter which plugs into the ratemeter and uses the same detection circuitry. On its own, the ratemeter is a compact, lightweight instrument which can be used to determine the approximate level of radioactivity being emitted from a source. This is adequate for most practical measurements, but where small changes in background radiation are being compared the accuracy provided by a meter display may not be good enough. Using the counter as well, it is possible to record the total number of pulses detected over long periods of time and so make accurate comparisons.



The ratemeter has three count ranges, from 18mR/h (milliRem/hour) FSD up to 2.5 R/h (Rem/hour) FSD. A loudspeaker is provided to give an audible indication of count rate (the traditional 'click-click' sound). The detector is a Geiger-Muller tube which is housed in a small plastic case and which attaches to the ratemeter by means of a length of cable and a plug. The unit is powered by four 1.5V cells and the circuit has been carefully designed to ensure as low a current consumption as possible, allowing the ratemeter to be used 'in the field' for long periods of time.

The separate counter has a four-digit LCD display and indicates the number of pulses detected by the ratemeter within a given time. No timing circuitry was considered necessary — the user simply starts the count when ready and then times it using a watch or timer. When the timing period is complete, the counter is stopped and the total number of pulses noted. A useful feature is the halt switch which freezes the display but allows the count to continue undisturbed. In this way, the count can be taken over a period of, say, ten minutes while the progress of the count can be noted at intervals of one minute or whatever by halting the display long enough to note the count figure down. The counter draws its power from the ratemeter using the same lead that transfers the pulse information, and even with the two units in operation the battery consumption is low enough to allow long periods of uninterrupted use.

The complete instrument costs under £100 to build (including the Geiger-Muller tube) and uses readily-available components. It is simple to calibrate, straightforward in operation and ideal for use in a wide range of applications from teaching physics through to checking food and drink.

The Geiger-Muller Tube

The first decision to be made when designing a piece of radiation monitoring equipment is the type of detector to use. Most detectors rely on the fact that radiation causes ionization when it interacts with matter, removing electons from atoms and so creating charged particles. The ions will be attracted to an electric field placed in their path and the resulting fluctuation in the strength of the field can be detected and displayed. This is the principle on which ionisation chambers work, but the output signals they produce are very small.

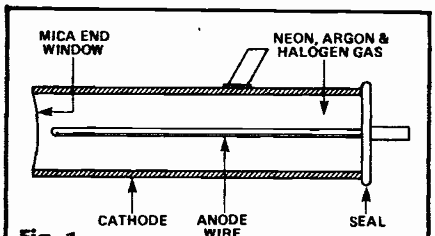


Fig. 1

If the strength of the electric field is increased, the ions accelerated by it gain enough energy to ionise more gas atoms in collisions. This results in an avalanche of electrons and so-called gas amplification of the original ion pulse. This effect is used in the Geiger-Muller tube (Fig. 1). It consists of a wire anode concentric with a metal cylinder and is filled with gas at less than atmospheric pressure. When the tube is fed from a high voltage supply, a strong field will be present around the anode wire and avalanching will take place whenever ions are present. Ultraviolet photons are created which cause the avalanche effect to spread by the photoelectric emission of electons. The result is that the tube presents a low impedance to the high voltage supply and pulls the anode voltage down, creating a large, negative-going pulse which is easy to couple into electronic circuitry.

One problem with this arrangement is that, once initiated, the avalanche may be sustained by slower positive ions which reach the cathode and knock out secondary electrons. This is overcome by means of a 'quenching' agent which absorbs the energy from the positive ions and so prevents further avalanching. In the Geiger-Muller tube specified for this project, the Mullard ZP1400, the quenching agent is a halogen gas.

The mica end window is very thin and bows inward due to the reduced pressure inside the tube. It allows beta radiation to enter the tube and, if sufficiently thin, will also admit alpha particles. Special tubes with a thinner window can be used when alpha particles must be detected. An example of this type is the Mullard ZP1401. Gamma rays are detected in the gas and by scattering electrons out of the cathode material into the gas. Other ionising radiations may also be detected.

The Circuit

The GM tube requires a 500V supply for operation and this is generated by a switching power supply from the 6V battery supply. The current drawn by the tube is only a few microamps so very little power needs to be generated. To minimise battery drain, the transformer is driven with short pulses at long intervals, causing it to generate a characteristic exponentially-decaying 'ring'. The stage using Q1 and Q2 (Fig. 2) is an astable pulse generator. Q3 amplifies and further shortens the pulse and Q4 drives the transformer.

The stepped up output voltage from the transformer is further increased by the voltage doubling circuit, developing the bias for the GM Tube.

Radiation causes output pulses to appear across the GM Tube when the gas around the anode becomes ionised. The pulse has a rapid rise and a slower decay caused by the stray capacitance in the circuit. Q5 squares the pulse and Q6 further squares and amplifies the signal. The output of Q6 drives the display and output circuits and is also coupled to Q4, the HT driver stage. This feedback causes each GM pulse to provide additional HT drive. Consequently, the output of the GM bias supply is maintained when the GM tube current drain increases due to ionisation by radiation. This method has been used successfully in other designs without any oscillator, relying on the background counts to drive the HT generator. However, if there were no charge on the HT capacitors, the supply would have to be primed to start it up. Hence the provision of the 'idling' oscillator, Q1, Q2.

Each pulse is made audible by Q7 which drives a short current pulse through a miniature loudspeaker. The clicking gives a satisfyingly 'authentic' direct indication that the GM tube is detecting background radiation, and an immediate indication when radiation increases. The current amplifier Q8, Q9 charges capacitors for a ratemeter display. Each pulse adds to the charge on a capacitor which leaks away through the indicating meter. The meter gives a reading proportional to the rate of the pulses and is switched to give ranges of different

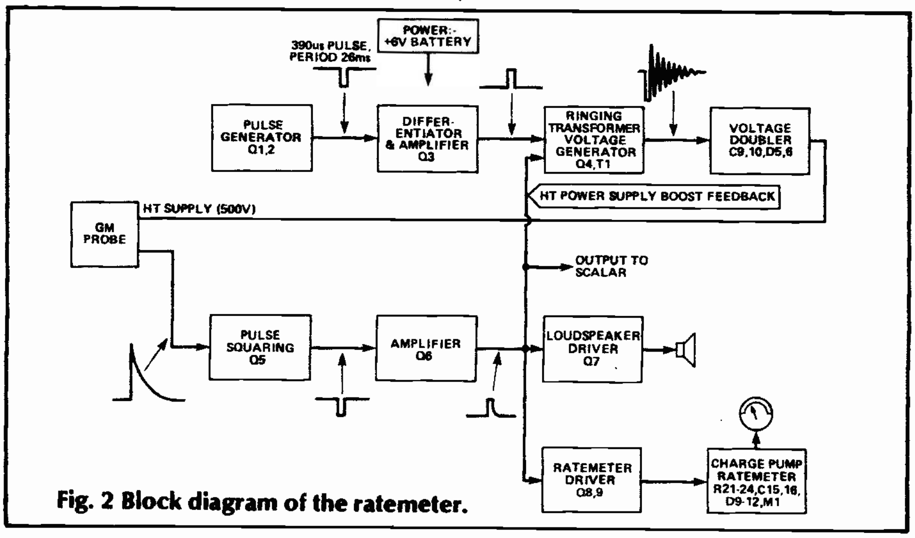


Fig. 2 Block diagram of the ratemeter.

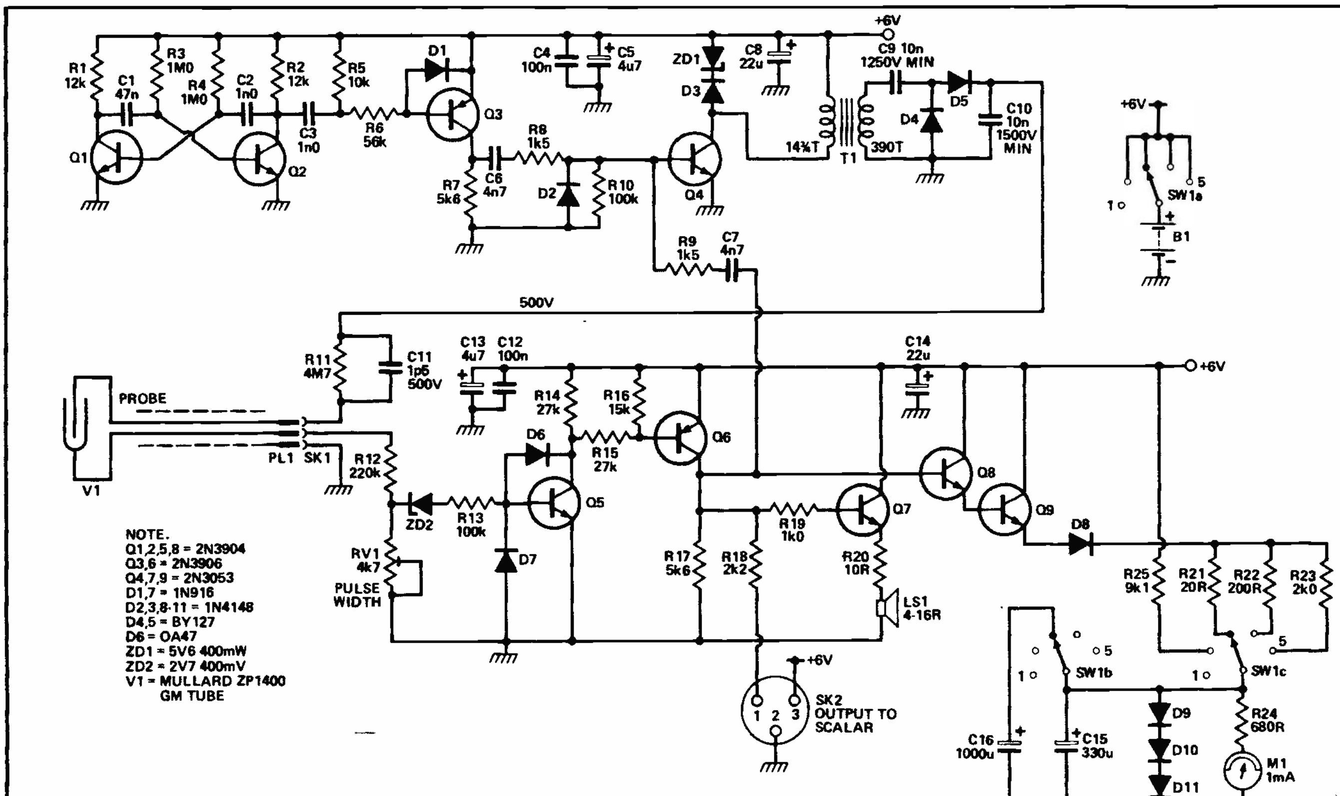


Fig. 3 Circuit diagram of the ratemeter.

HOW IT WORKS

No ICs are used in the detector unit as there is not much need for them and discrete transistors are more resistant to damage by radiation than ICs — although we hope that the user does not encounter such levels of radiation!

The range switch SW1 also acts as an on-off switch for the battery. Due to the low current drain, the unit can be left on to monitor for long periods. A further position on the switch is used for battery checking.

Q1 and Q2 form a conventional astable multivibrator with asymmetrical time periods. C3 shortens the pulse which is squared and amplified by Q3 while D1 prevent Vbe breakdown. C6 and R8 shorten the pulse further to about 25 us. Q4 drives the primary of T1, the current being limited mainly by the inductance of the primary. To provide sufficient inductance, a pot core with a high Al was used in the prototype (Mullard FX2241). Due to the large step-up ratio, about 250 V peak-to-peak is developed in the secondary and appears across C9 and D4. Voltage stabilisation is achieved by the flyback limiting action of Zener diode ZD1 and rectifier D3.

Voltage doubling occurs as the voltage on C9 is added to that rectified by D5, so 500V appears across C10. These capacitors are generously rated for reliability and must have very low leakage. R11 limits the GM tube current as specified by the data sheet, and C11 helps to shorten the decay of the pulse. The decay time of the pulse is also dependent on the capacitance of the cable connecting the GM tube to SK1. R12 and RV1 attenuate the 100 V or so peak pulse and set the level for the pulse-squaring stage, Q5, reducing the pulse length to 25 us. D7 prevents Vbe breakdown and D6 prevents Vcb breakdown in the vicinity of the 100 V pulse. D6 also limits charge storage which is why it must be a Germanium or Schottky type. ZD2 raises the threshold voltage to a convenient level.

Q6 further amplifies the pulse and enables the distribution of a CMOS-compatible signal to the external socket, the HT feedback input at C7, the loudspeaker driver Q7 and the meter driver Q8/9. D8 prevents Vbe breakdown in Q9.

R21–23 are switched to change meter sensitivity. R24 plus the meter resistance (300 ohms) should be a total of 1k0 (to within about 2%) to make a 1V FSD meter at 1 mA full load. R24 would have to be changed if a meter with a different resistance were used. D9, 10 and 11 protect the mteter from surges. C16 provides additional smoothing capacity on the lowest count rate range.

sensitivity. The capacitance helps to smooth out the variations caused by the random arrival of pulses, especially at low levels.

The pulse is also made available to external circuitry, such as the scalar/counter, by means of an output socket which also carries the supply voltage.

The HT supply and output circuits take large current pulses from the supply, but the duty cycle is very low over the normal range of operation so the standing current drain is also very low. Capacitive decoupling of the supply prevents malfunction due to the current spikes. The current drain is 0.62 mA with normal background radiation, rising to a few mA on exposure to radiation sources.

Construction

The majority of the components are mounted on two PCBs, a main board which carries the bulk of the circuitry and a smaller board which carries the ranging circuitry and bolts directly onto the back of the meter terminals. The remainder of the components are attached directly or indirectly to the front panel, with the exception of the battery pack which is secured in the bottom of the case.

The PCBs are double-sided with tracks on one side and a ground plane on the component side. The use of a ground plane is highly recommended for the main board because this helps to isolate the HT supply pulses from the GM tube pulse amplifier and reduces radiation of RF interference from the HT supply. It also helps to reduce sensitivity to spurious pickup from mobile radio transmitters and other sources which might interfere with the readings. The main board should be of fibreglass, as any leakage current is liable to excessively load the HT supply.

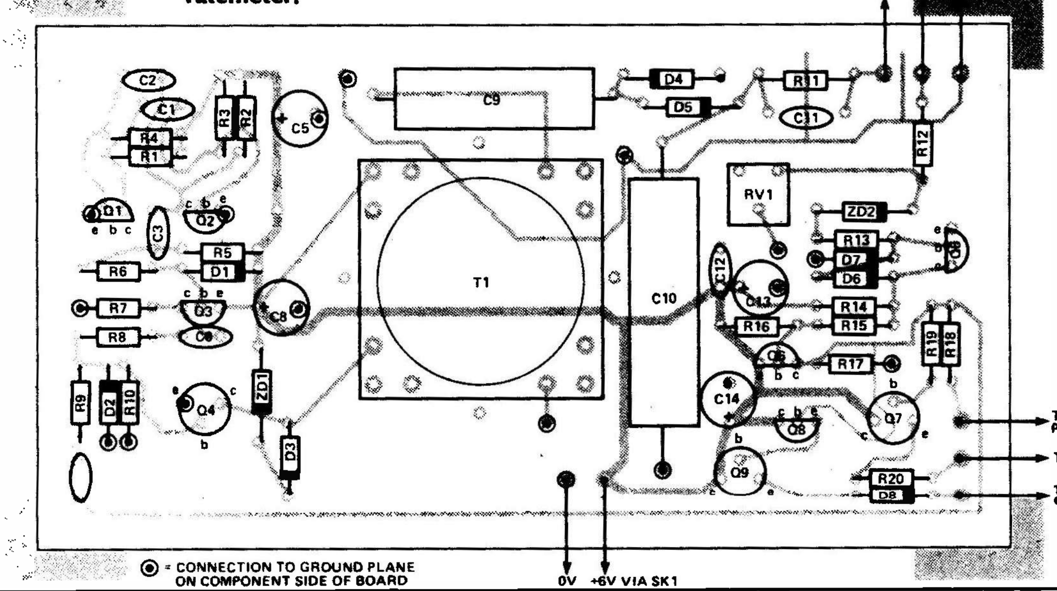
The ground plane on the prototypes was created simply by leaving the reverse side of the boards un-etched. This removed the need to make up a set of ground-plane foil patterns. Boards supplied by our PCB Service will be made in a similar way, so the component lead holes must be isolated from the ground plane using a counterbore tool or drill, and plenty of clearance to the ground plane should be left around the 500V HT circuitry.

None of the components used present any handling problems so there is no need for any precautions. Begin by soldering into place the passive components (resistors, capacitors, etc) then move on to the diodes and finally the transistors. Take the usual care with the semiconductors and the tantalum capacitors, all of which must be installed the right way around if the circuit is to work correctly. Note that there is no 0V rail on the track side of the main board and that grounded leads must be soldered to the ground plane on the top side of the board.

The coil assembly, T1, cannot be completely assembled at this stage because its turns ratio must be adjusted as part of the calibration procedure. The secondary is wound first and consists of about 390 turns of 42–48 SWG enamelled copper wire. This should be wound in successive layers of 50–60 turns with an insulating sheet of paper or thin plastic tape between each layer. An initial primary winding of 30–38 SWG enamelled copper wire is then added, consisting of as many turns as will fit on the bobbin in one layer. The number of turns in this winding will later be adjusted to give the correct output voltage. Temporarily fit the coil assembly onto the PCB and solder the leads into place.

The prototype was constructed in a plastic box with an aluminium front panel. The size of the box was determined largely by the size of several key components,

Fig. 4 Component overlay for the main board of the ratemeter.



PARTS LIST — MAIN BOARD

RESISTORS

R1, 2	12k
R3, 4	1M0
R5	10k
R6	56k
R7, 17	5k6
R8, 9	1k5
R10, 13	100k
R11	4M7 0.5W
R12	220k
R14, 15	27k
R16	15k
R18	2k2
R19	1k0
R20	10R
RV1	4k7 miniature enclosed horizontal preset

CAPACITORS

C1	47n ceramic
C2, 3	1n0 ceramic
C4, 12	100n ceramic
C5 13	4u7 10V tantalum
C6, 7	4n7 ceramic
C8, 14	22u 10V tantalum
C9	10n 1250V (minimum) polyester, polycarbonate, polystyrene or mylar
C10	10n 1500V (minimum) polyester, polycarbonate, polystyrene or mylar
C11	1p5 500V (minimum) ceramic

SEMICONDUCTORS

Q1, 2, 5, 8	2N3904
Q3, 6	2N3906
Q4, 7, 9	2N3053
D1, 7	1N916
D2, 3, 8	1N4148
D4, 5	BY127
D6	0A47
ZD1	5V6 400mW
ZD2	2V7 400mW

MISCELLANEOUS

T1	Mullard FX 2241, core grade A13 (minimum Al = 5815) with clips and tagboard. No trimming core.

PCB; 30-38 SWG and 42-48 SWG enamelled copper wire for T1.

notably the meter and the loudspeaker which were both purchased ex-equipment in order to keep costs down. Constructors using different sizes of meter and loudspeaker may prefer to use a larger or smaller case. The main PCB was designed so that it would fit neatly into the guide slots of the prototype case, but there is no reason why it should not be mounted on stand-off pillars should a different type of case be used.

The meter, the switch, the loudspeaker, the GM tube socket and the count-output socket were all mounted on the front panel, but it may be preferable to mount some of these components (the two sockets, for example) on the sides of the case. Some sort of handle will also be required, and it is well worth experimenting with the position of this and of the other major components in order to acheive the best distribution of the weight. The complete unit is very light, but it will be far more comfortable to use for long periods if it is well balanced.

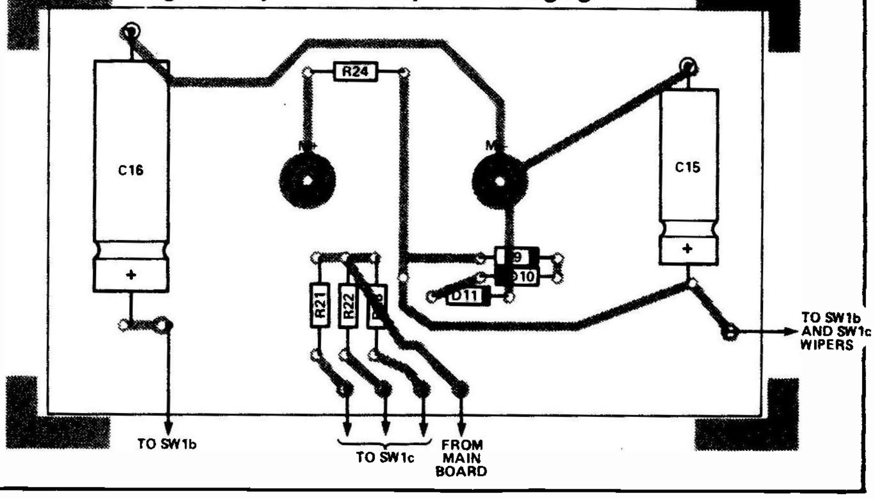
The interwiring of the major components is shown in the overlay diagrams. The only difficult item is the switch which has quite a number of wires running to and from it and must also have some of its contacts linked. Note that R25 is mounted between two of the switch contacts. The 6V battery pack can be mounted in the bottom of the case by any convenient means. In the prototype, it was bolted to a spare piece of copper-clad board which was cut so that it would fit into the PCB guide slots.

The probe should be housed in a tubular plastic container with a detachable end cap. The dimensions aren't critical and it should be possible to find something in the kitchen or the medicine cabinet that will do the job. A hole is needed in the bottom of the container to carry the screened cable and a larger hole should be cut in the cap to enable beta particles (and alpha particles if you are using the ZP1401) to reach the GM tube. The lower hole should be fitted with a rubber grommet while the upper hole must be covered with a wire mesh (the type used in tea strainers is ideal) in order to project the fragile end window of the tube.

A new GM tube should be purchased to eliminate any doubts about its efficiency. It will come fitted with a protective plastic cap which should be left on until it has been wired to the cable, and drawn into the container cushioned in a foam plastic wrapping. The anode pin is connected via a push-on clip which is supplied with the tube. Take care not to touch the end window with anything when you fit the cap.

Ideally the GM tube should be mounted in low background count materials, but the plastic materials used were found to make no significant difference to background count. The probe assembly should be kept clean and there must be no solder splashes. The box that the GM tube comes in may be adapted with its foam packing as a storage box for the finished probe.

Fig. 5 Component overlay for the ranging board.



PARTS LIST — RANGING BOARD AND CASE

RESISTORS			
R21	20R 1%		
R22	200R 1%		
R23	2k0 1%		
R24	680R (or 1k0 minus meter resistance)		
R15	9k1 5%		
CAPACITORS			
C15	330u 10V tantalum		
C16	100u 10V tantalum		
SEMICONDUCTORS			
D9, 10, 11	1N4148		
MISCELLANEOUS			
B1	4 x 1.5V AA cells in holder		
LS1	4-16R miniature loudspeaker		
M1	1mA meter movement		
PL1	¼" stereo jack plug		
SK1	¼" stereo jack socket		
SK2	5-pin 180° DIN socket		
SW1	3-pole 5-way rotary switch (or 4-pole 6-way MAKA switch with adjustable end stop, one pole unused)		
V1	Mullard ZP1400 Geiger-Muller tube (or ZP1401 — see text)		

PCB; case; handle; knob for SW1; 2-core screened cable, plastic tube with cap, fine wire mesh, mounting foam and rubber grommet, all for probe; connecting wire, nuts, bolts, etc.

Calibration

The GM tube supply voltage should be 500V ±25V, as measured by a 5% instrument. This is achieved by adjusting the primary turns of T1. Connect the ratemeter to its battery pack and switch on. If the circuit waveforms are present as shown in Fig. 2, the HT voltage can be checked. This will require an oscilloscope with a 1V/cm range, 1Mohm input, and a 10Mohm x 10 probe. A further x 10 probe must be made from two 33 Mohm and one 22 Mohm resistors in series, or one 90 Mohm 500V resistor. The high impedance enables the oscilloscope to display the voltage when the probe is touched to the HT supply. Even so, because of the low current available the HT capacitor soon discharges, so the probe must be briefly tapped on the HT output while observing the oscilloscope. The effective multiplication is 100 times, so a 5 cm deflection is required. If the HT voltage is out of limits, the primary winding must be rewound. The number of turns required will be equal to:

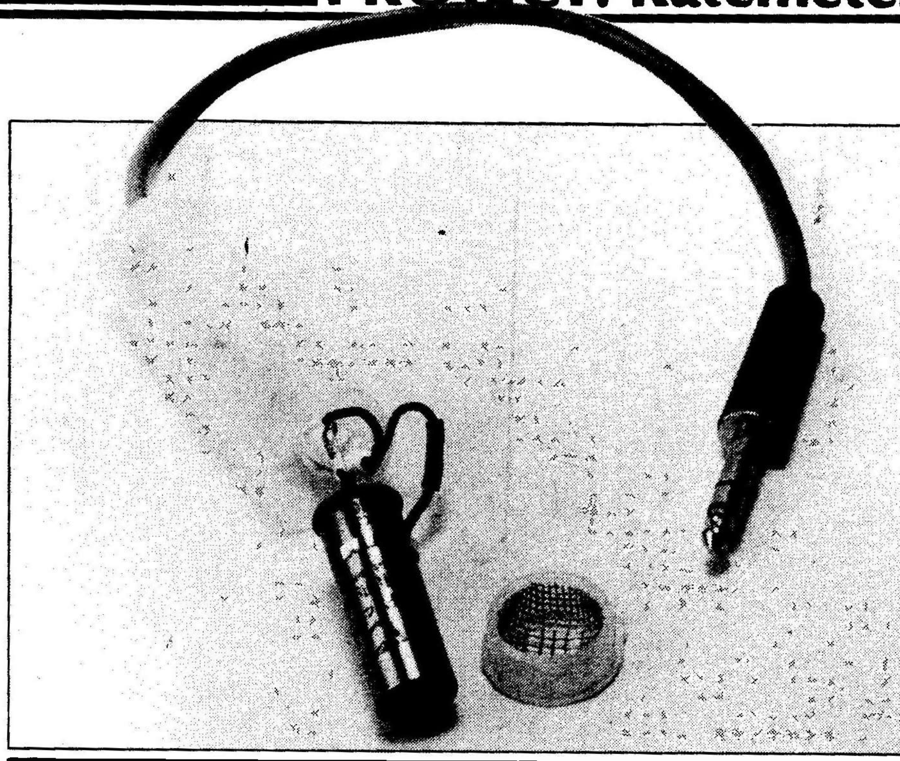
$$\text{original HT voltage} \times \frac{\text{original no. of primary turns}}{500}$$

The primary of T1 should be rewound to the correct value and replaced on the PCB. Check the voltage again to make sure that it falls within the specified limits, then secure the coil assembly to the PCB using cable ties passed through the holes provided.

Next, the trimmer RV1 should be adjusted so that the output pulse at Q6 collector is 25 us wide at 50% peak to peak. This standardises against variations in GM tube load capacitance.

Now the ratemeter calibration can be considered. The prototype scale (Fig. 7) was calibrated by setting various count rates using an object with radium paint, and referring the count rate to the dose rate graph given in the GM tube data sheet. A range of points were plotted for range 1, giving a curved graph. The output current characteristic calculated for the output circuit showed a similar curve. The non-linearity (due to non constant charging current) can be fitted to the measurements by establishing a highest reading on each scale, and multiplying the rate by a scale factor for each other current value. On range 2 and 3, extrapolation was necessary due to the undesirability of obtaining more powerful sources. Note that the tube specification is limited to 1 R/hr, whereas Range 3 scale ends at 2.5 R/hr.

The method will undoubtedly include sizeable errors. More accurate calibration for a particular counter and probe will require a precise radiation dosimetry set-up with calibrated standard radiation sources. This will only be available to a few specialists. Also, sensitivity can be expected to vary depending on the type of radiation and its energy. However, if the user's main concern is to minimise exposure to radiation then measurements are likely to be relative anyway. If a piece of meat has twice the background level of radiation then you would probably not want to eat it regardless of what the measured level in Rems is. In the same way, many radioactivity experiments can be done using count rates as relative quantities.



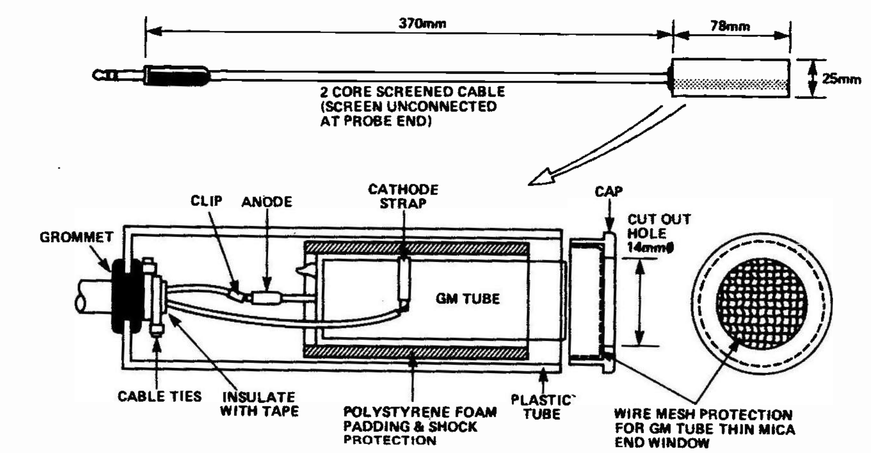


Fig. 6 Construction of the Geiger-Muller tube probe.

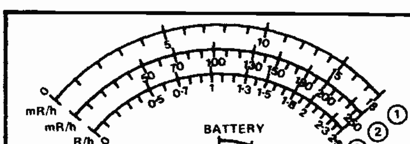


Fig. 7 Calibration guide for the meter scale.

The accompanying counter unit will be described next month, along with some notes on its operation and on the use of the two instruments.

ETI

BUYLINES

The resistors, the semiconductors and most of the capacitors are popular types and can be obtained from any number of our regular advertisers and other component suppliers. High voltage capacitors are available from STC Electronic Services (Tel (0279) 441 687), Electrovalue (Tel (0784) 33 603) and MS Components (01-670 4466) among others. Many of the types they stock offer far higher voltage ratings than are needed here, but make sure that the type you buy will fit on the PCB! STC also stock the FX2241 pot core and we suspect that Electrovalue will have an equivalent somewhere among their vast range. Almost any 1mA panel meter should do for M1, but if you want to use the PCB layout without modifications you must choose a type with 1" (25.4mm) spacing between the terminals. The ZP1400 Geiger-Muller tube (which is sensistive to beta and gamma radiation) and the ZP1401 tube (which also senses alpha particles) can both be obtained from Alrad Instruments Ltd, Turnpike Road Industrial Estate, Newbury RG13 2NS, tel (0635) 30345. As a guide, the latest prices we have for these two tubes are £37.50 and £40.50 respectively plus VAT, but you should check before ordering in case these have changed. The PCBs are available from our PCB Service — see page 56.

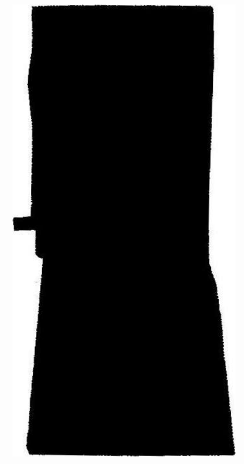
PHOTO PROCESS CONTROLLER

Okay, so it's just an an enlarger timer, but Ian Coughlan's design is the enlarger timer to end all others.

Printing your own photographs is one of the most pleasant aspects of photography. Unfortunately, as with most enjoyable pastimes, there are interesting bits and there are boring bits. One of the more excruciatingly boring bits is timing the exposure when making prints. Peering at a stop clock under safelight for many tens of seconds is no-one's idea of fun. When producing colour prints, it becomes impossible to see the stop-clock, since colour papers cannot practicably be handled under safelight conditions. In any case, colour or monochrome, the safelight ought to be off when making the exposure.

This Photo Process Controller is a neat solution to the problem. Set the desired exposure time on a pair of rotary switches, press the EXPOSE button, and the enlarger is switched on for precisely the desired time. The exposure time altered can be set anywhere between one and 99 seconds, in one second intervals. In addition to this basic function, the controller has a number of advanced features.

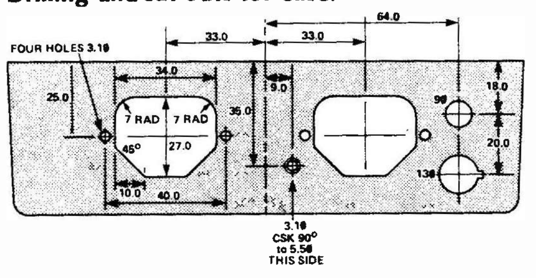
Pressing the FOCUS button will switch the enlarger on, without the timing function, so the enlarger can be set for size, composition and focus. Pressing the RESET button will switch the enlarger off again. The RESET button can also be used to abort the timing function whenever needed. In addition to the power-outlet socket for the enlarger, another is provided for the safelight. Whenever the enlarger is on, the safelight is off. There are two reasons for this. The first is it

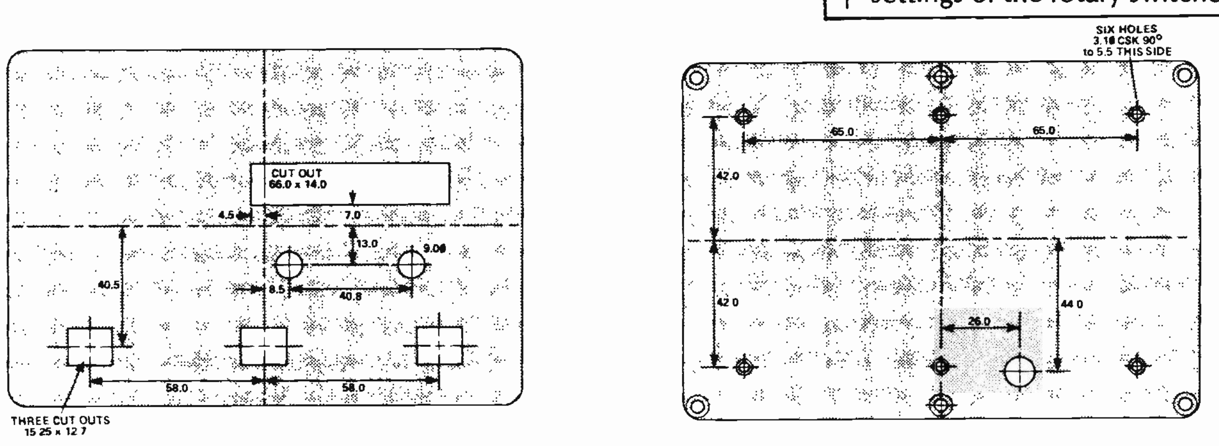


makes focusing much easier. The second is that, although photographic papers are relatively insensitive to the correct safelight, they are not totally insensitive. If a negative requires a long exposure time then switching off the safelight reduces the chances of the paper being 'fogged'. Two LEDs show the status of the power to the enlarger and safelight.

The exposure time set on the rotary switches is shown on a pair of red 7-segment displays. When the paper is being exposed, the displays show the exposure time remaining. When the time is over, the displays revert to showing the settings of the rotary switches.

Fig. 1 Drilling and cut outs for case.





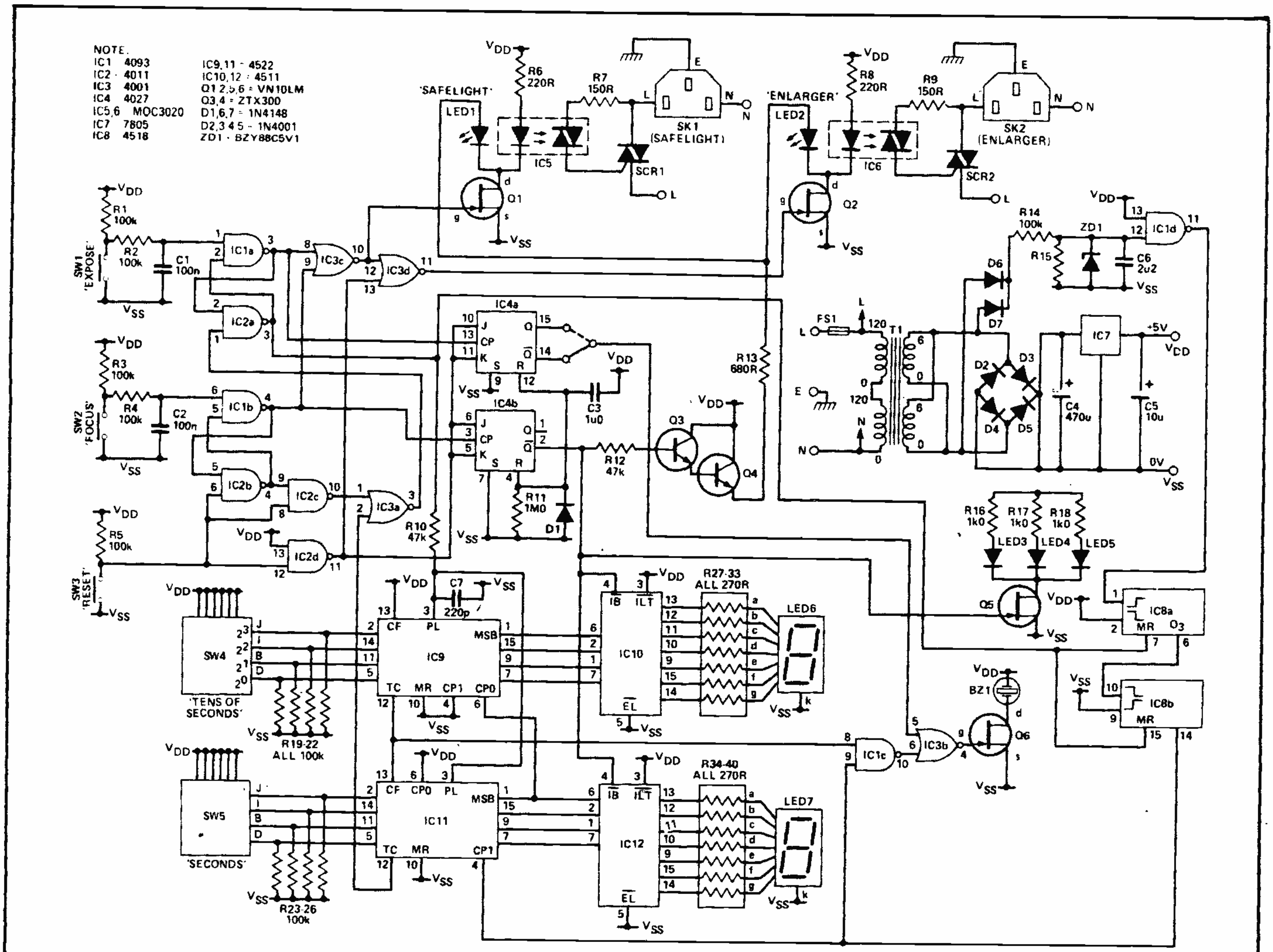


Fig. 2 Circuit diagram of the Photo Process Controller.

BUYLINES

The resitor networks are from RS (stock no. 140–029) and are available through the Electromail service (0563) 204555. Alternatively eight individual 270R resistors may be used in place of each network. The opto-isolator triacs are General Electric H11J1 and are available from RS as 308–196, as are the VN10LM VMOS transistors (RS 295–107). The 7 segment displays (LED 6, 7) are RS 587–917 and the buzzer is RS 249–429. The rotary BCD switches are RS 327–939 and the mains transformer is RS 207–829 although other models would be adequate. The box used for the prototype was also from RS (509–995 but, again, others conforming to the measurements given would serve. The collet knobs should be for a 6mm shaft if the RS BCD switches are used. The push button switches in the prototype are from Roxborough Electronics (no. 8125–J85) but others may be used instead. Note the integral LED in these switches should be red. Depending on the box and LEDs used the T0–92 transistor sockets and the DIL turned pin sockets may not be required to raise the LEDs. Any suitably sized display bezel can be used.

The three pushbuttons are illuminated by a further three red LEDs, making them easy to find in the dark. Since monochrome photographic papers are generally insensitive to red light, all these LEDs will not produce a significant fogging effect — so long as the paper is kept at a reasonable distance from the Timer, say a couple of feet or so. However, it may be that it is desirable not to have these LEDs illuminated, such as when printing on colour paper, which is sensitive to red light. The LEDs — including the displays — may be switched off or on as you please by holding down the RESET button and pressing the FOCUS button.

As if this wasn't enough, an audible bleep is produced for each of the last ten seconds of the exposure time. If the bleep becomes irritating, that too can be enabled or disabled by pressing the EXPOSE button with the RESET button held down.

The Controller has been designed to be easy to use even in total darkness, with all the LEDs and displays off. The three pushbuttons are quite large and widely spaced; and the rotary switches have end-stops, so they can be operated by feel, counting the clicks.

Construction

The prototype was designed around a 171 x 121 x 55 mm die-cast box. If using the foil patterns given, it is recommended you stick with this box size. Drill the box and base as shown in Fig. 2. The rectangular cut-out on the top of the box is for the display bezel. This can be to suit the bezel used. Similarly, the holes for the pushbuttons must be cut or drilled to suit whatever type is used. The box can then be painted and lettered if desired.

Fix the mains connectors and the fuse-holder to the rear of the box. Fix a grommet suitable for 3-core 3-amp mains cable. The earthing screw should be firmly secured to the box, and then four solder-tags attached to the screw. The pushbuttons are a push fit into the box. Fix another grommet to the base, and stick on four plastic

HOW IT WORKS

The timing reference for the controller is the mains supply. Note this design cannot be used on mains supplies with a frequency other than 50Hz.

The 50Hz is doubled by full-wave rectification, clamped by ZD1 to 5 volts, and filtered by R14, R15 , and C6 to remove mains-borne interference, before emerging from IC1 pin 11 as a 100Hz square-wave. This is then divided by a pair of decade counters, IC8, to give one pulse per second.

The two rotary switches SW4 and SW5, produce logic levels that depend on their settings. So the binary codes decimal (BCD) equivalents of their settings can be placed on the inputs of a pair of programmable down-counters, IC9 and IC11. As long as the Controller is not in EXPOSE mode, the Parallel Load (PL) inputs to IC9 and IC11 will be at logic 1, and hence the outputs of these down-counters (and the displays, via decoder-drivers IC10 and IC12) will mimic the settings of the rotary switches.

When the EXPOSE button is pressed, the control circuitry will put a logic 0 on the PL inputs, latching the settings of the rotary switches into the counters. Altering the settings of the switches will not affect the counters of the displays. One second later, the first of the clock-pulses will arrive from IC8, causing IC11 — the 'seconds' counter — to decrement its count. The next clock pulse will decrement it again, and so on, until 0 is reached. Assuming the 'tens-of-seconds' counter — IC9 — is not also at 0, the next clock-pulse will change the seconds counter to 9, and the tens-of-seconds counter will decrement, since it is clocked from the most significant bit of the seconds counter. While this is happening, the displays are changing with the counters, showing the time remaining.

Both IC9 and IC11 have a Terminal Count (TC) output, which goes to logic 1 when the counter reaches 0 as long as the Carry Forward (CF) input of that counter is also at a logic 1.

The CF Input of the tens-of-seconds counter is tied to Vdd, and so the TC output of that counter will go to logic 1 as soon as 0 is reached. The CF input of the seconds counter, however, depends on the TC output of the tens-of-seconds counter, and so the TC output of the seconds counter will not go to logic 1 unless both counters are at 0, signifying the end of the exposure-time. That logic 1 is passed to the control circuitry which resets the controller and puts a logic 1 on the PL inputs of both counters causing them and the displays, to revert to the settings of the rotary switches. The PL inputs must not go to logic 1 too soon after the TC output has gone to logic 1, or the TC outputs tend to latch-up. R10 and C7 serve to delay slightly the logic 1 from the control circuitry.

In describing the operation of the control circuitry we start with the Enlarger Timer reset with the enlarger off and the safelight on.

IC1a and IC2a form a latch circuit. IC1 pin 3 will be at a logic 0, and IC2 pin 3 will be at logic 1. IC1b and IC2b form an identical latch circuit with IC1 pin 4 at logic 0 and IC2 pin 4 at logic 1. So IC3 pins 8 and 9 will see a pair of logic 0s making IC3 pin 10 go to logic 1 and turning on Q1. This will light LED1 and activate the optically-coupled triac, IC5 which, in turn, is used to switch a much larger triac, SCR1 to control the safelight.

The logic 1 from IC3 pin 10 is inverted by IC3d, turning off Q2, ICL6, SCR2 and the enlarger.

When the FOCUS button is pressed, IC1 pin 6 will be pulled to logic 0, setting the latch formed by IC1b and IC2b. IC1 pin 4 will go to logic 1, and IC2 pin 4 will go to logic 0. The cross-connection of outputs to inputs will ensure that the latch will remain in this state until the RESET button is pressed. The logic 1 from IC1 pin 4 will be inverted by IC3c (turning the safelight off), and re-inverted by IC3d (turning the enlarger on).

When the RESET button is pressed, the latch will revert to its previous state, and so the enlarger will go off, and the safelight will go on.

When the EXPOSE button is pressed, the result is very similar to that obtained when the FOCUS button was pressed, except now the timing function has to be enabled. IC2 pin 3 will go a logic 0, removing the reset from IC8 pins 7 and 15, allowing the decade counters to start dividing the 100Hz clock frequency down to 1Hz. At the same time, the PL inputs to IC9 and IC11 go to logic 0, locking-out the parallel inputs from the rotary switches, and allowing the counters (and the displays) to step downwards.

When both counters have reached 0, the TC output of IC11 will go to logic 1, and this is inverted by IC3a to produce a logic 0 to reset the latch (IC1a and IC2a) which in turn resets IC8, and puts a slightly delayed logic 1 on the PL inputs to IC9 and IC11, so they revert to the settings of the rotary switches.

The timing function can be interrupted before the exposure-time is over, by pressing the RESET button. This will reset the latch in the same way as the two counters reaching 0, as described above.

The displays are enabled by a logic 1 on pin 4 of the decoder-drivers (IC10 and IC12). similarly, LED1 and LED2 are enabled by turning on the Darlington-pair Q3 and Q4. Finally, LED3, LED4, and LED5, are enabled by turning on Q5. Thus the status of all the LEDs will be controlled by the logic level on IC4 pin 2.

IC4 is a dual JK flip-flop. When power is first applied to the controller, both flip-flops will be reset, due to the effect of R11 and C3 on the reset pins of IC4. Pin 2 of IC4 will be at a logic 1, and the LEDs and the displays will be on. The Clock Pulse (CP) input of IC4b will go to a logic 1 each time the FOCUS button is pressed, but since the J and K inputs are both at logic 0, the flip-flop will not change states and the LEDs will remain on.

However, press and hold down the RESET button, and IC2 pin 11 will go to logic 1, as will the J and K inputs. When the FOCUS button is now pressed, the flip-flop will toggle, turning the LEDs on or off as desired.

The CP requires a single pulse each time the FOCUS button is pressed, so the contact-bounce of the button is eliminated by R3, R4, and C2. IC3d prevents the enlarger switching on when both the RESET and FOCUS buttons are pressed.

Remember the TC output of the tens-of-seconds counter (IC9) will be at logic 1 when the contents of that counter are at 0. This logic 1 is used to gate the 1Hz pulses from IC8, so IC1 pin 10 will have a negative-going pulse for each of the remaining ten seconds of the exposure-time. As long as IC3 pin 5 is at logic 0, these pulses will appear inverted at IC3 pin 4, turning on Q6 and the buzzer when positive. If IC3 pin 5 is taken to logic 1, then IC3 pin 4 will always be at logic 0, and the buzzer will not sound.

IC3 pin 5 is controlled by the other JK flip-flop in IC4. This flip-flop functions in exactly the same way as the one controlling the display status, except that in this case, the clock pulse is provided by pressing the EXPOSE button.

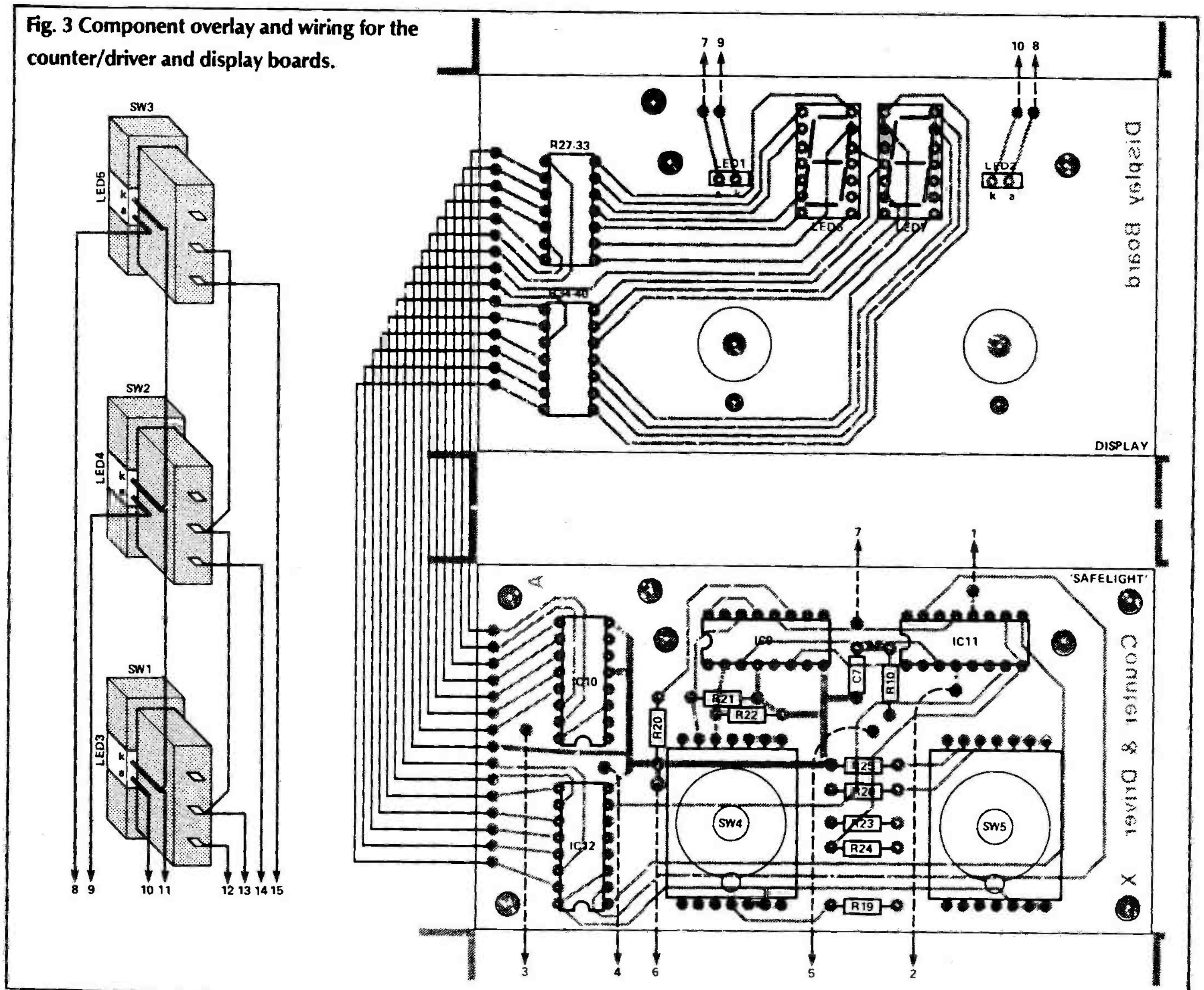
This flip-flop will also be reset on power-up, although the constructor can decide whether he wants the buzzer enabled or disabled. In the former case, IC3 pin 5 should be connected to IC4 pin 15, and in the latter case IC3 pin 5 should be connected to IC4 pin 14. A link on the PCB provides for this choice.

feet. These are necessary to provide a gap for the sound of the buzzer to get out.

Using 3 amp mains connecting wire for the high-voltage parts, and any thin insulated connecting wire for the low-voltage parts, connect the case-mounted components as shown in Figs. 3 and 5. Use sleeving wherever possible. Put the box to one side.

Fit veropins to the three printed-circuit boards. The veropins should be inserted from the foil side of the large PCB, and from the component side of the other two PCBs. Fit the resistors and wire links first, then the capacitors, transistors, diodes, and the transformer. Note that DIL sockets can be used on both of the single-sided PCBs, but not on the double-sided counter/driver PCB as the IC legs are used as feed-throughs, and must be soldered on both sides. Wherever there is a pad on the component

Fig. 3 Component overlay and wiring for the counter/driver and display boards.



side of the double-sided PCB, this must be soldered on both sides. Use turned-pin DIL sockets for the two LED displays. These will lift the displays up by just the right amount. For the same reason, fit a small transistor mounting pad under each of the two LEDs next to the displays. It is not essential to use resistor networks on the display board — individual 270R ¼W resistors may be used, if preferred.

Connect flying-leads to each PCB and connect and fix the PCBs to each other. The central PCB mounts to the power/control PCB on four 12.7mm insulated stand-offs. The display PCB mounts to the counter/driver PCB on one 12.7mm insulated stand-off, and on the bushes of the rotary switches.

Do not force the boards together. If they don't fit easily, find out why. In particular, do not put any force on the rotary switches. If necessary, file the holes in the PCBs to a slightly larger diameter.

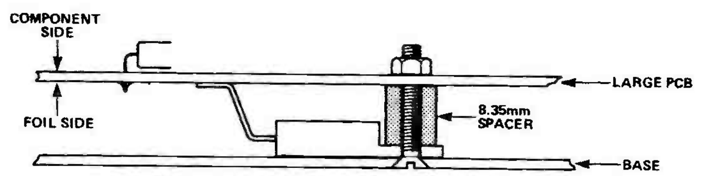


Fig. 4 Mounting for IC7 and SCR1, 2.

The 5V regulator and the two triacs mount in the base of the box, as shown in Fig. 4. Firstly, they must be soldered into place (on the underside of the PCB), so temporarily fix them into position on the PCB with a nut, screw, and an 8.75mm plastic spacer.

The Controller cannot be used to drive inductive loads, unless a snubber-network is connected across each of the triacs. If these are required, they should be soldered in place now, across MT1 and MT2 of each triac. Lay them flat under the PCB. A suitable snubber-network is RS 238-463. Remove the screws from the triacs and the 5V regulator, and carefully assemble the PCBs to the base. The complete assembly may now be offered up to the box, and the case-mounted components connected. Use cable ties to keep everything tidy. Assuming that the assembly fits into the box without any trouble, screw the base onto the box. Attach the two collet knobs and the Enlarger Timer is now complete and ready for testing.

Testing

Because mains voltages are present in the Controller, the importance of thorough testing cannot be too highly stressed.

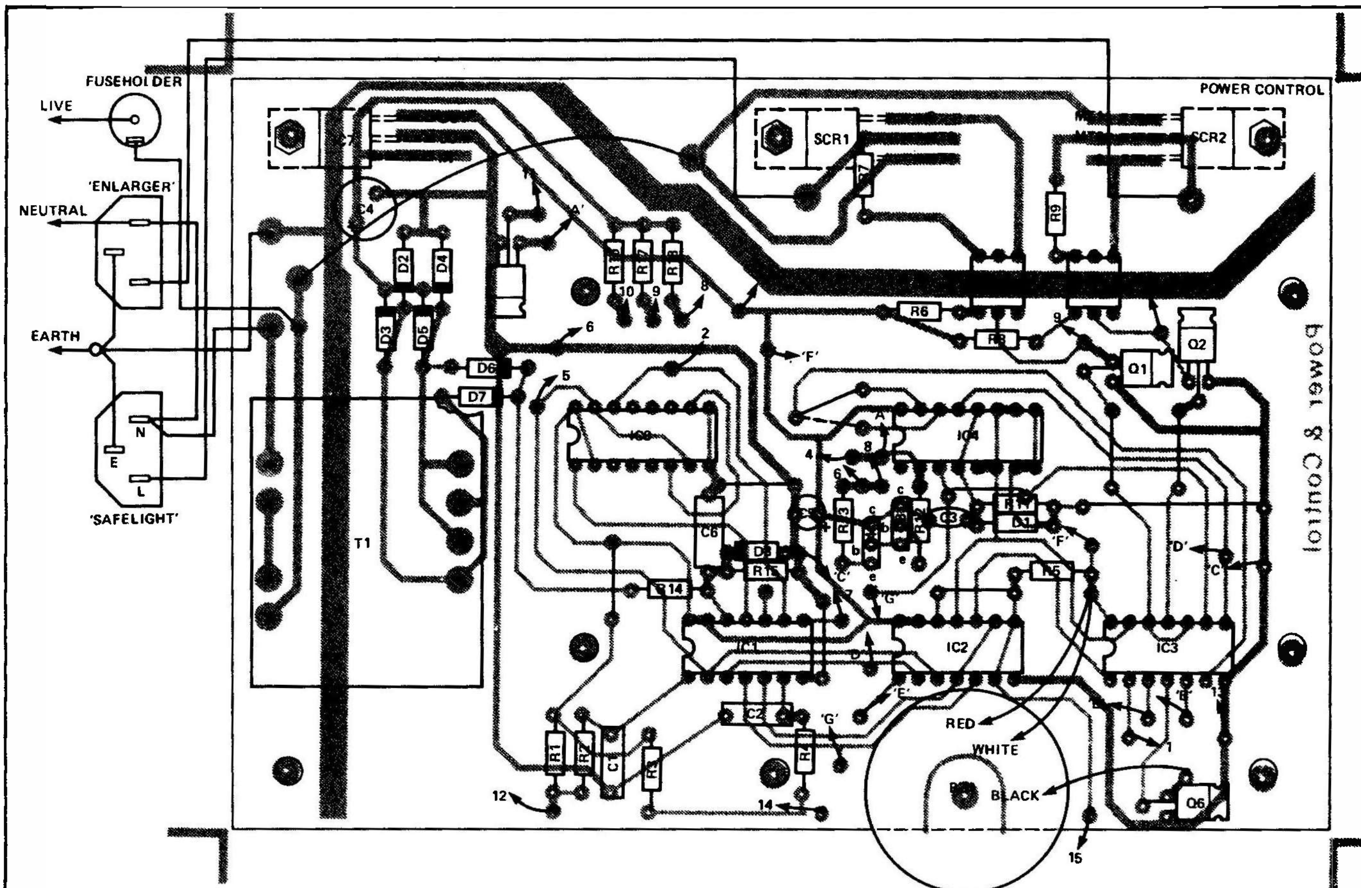


Fig. 5 Component overlay and wiring for the power/control board.

Using a multimeter on its lowest resistance range, check that the earth lead in the mains cable makes good contact with the metal box, and with the central pin of the two mains sockets. Fit a 1A fuse to the fuseholder, and with the multimeter set to its highest resistance range, check that there is good insulation between the earth wire and the live and neutral connections, including those in the mains sockets. Only when these checks have been made should mains be connected.

If all is well, the LEDs should light, and the display should show the settings of the rotary switches. If the safelight indicator is off, press the RESET button. Check that the Controller functions in the manner described. An ordinary 240 V lamp can be used to check the outputs of the mains connectors. If everything checks out OK, the Photo Process controller is complete and ready to use.

ETI

PARTS LIST

RESISTORS (all ¼W, 5%)	
R1, 2, 3, 4, 5, 14, 19, 20, 21, 22, 23, 24, 25, 26	100k
R6, 8	220R
R7, 9	150R
R10, 12	47k
R11	1M0
R13	680R
R15	470k
R16, 17, 18	1k0
R27–33, 34–40	DIL resistor network, seven individual resistors each 270R

CAPACITORS	
C1, 2	100n miniature layer
C3	1u0 35V tantalum bead
C4	470u 16V radial electrolytic
C5	10u 16V radial electrolytic
C6	2n2 polystyrene
C7	220p polystyrene

SEMICONDUCTORS	
IC1	4093
IC2	4011
IC3	4001
IC4	4027
IC5, 6	MOC3020
IC7	7805
IC8	4518
IC9, 11	4522
IC10, 12	4511
Q1, 2, 5, 6	VN10LM
Q3, 4	ZTX300
D1, 6, 7	1N4148
D2, 3, 4, 5	1N4001
ZD1	BZY88C5V1
LED1, 2	Rectangular red LED
LED6, 7	0.3" 7 segment LED display, common cathode
SCR1, 2	T2850D 8A 400V triac with insulated heat sink tab.

MISCELLANEOUS	
T1	3VA PCB-mount transformer 0–120, 0–120 primary 0–6, 0–6 secondary
SK1, 2	IEC Euro chassis mains outlet socket
SW1, 2, 3	Miniature push button switch (incl LED 3, 4, 5)
SW4, 5	PCB-mounted BCD rotary-switch
BZ1	5V Piezo sounder

PCB; die-cast box, 171x121x55mm; display bezel; collet knobs for 6mm shaft, and caps; cable grommets; fuse holder and 20mm 1A fuse; solder tags; silicone rubber sleeving; nylon spacers 8.75mm (3), 10.0mm (3), 12.7mm (5); DIL sockets, if required; T0–92 transistor pads (2); turned-pin DIL sockets 14-pin, for displays; Veropins; insulated connecting wire; screws, nuts, washers, etc; Plastic feet, paint, lettering, mains cable.

LOW-DISTORTION STEREO DECODER

John Linsley Hood brings you a practical FM receiver design which uses some of the techniques described in his recent theoretical series. To begin with here's the stereo decoder board.

The stereo decoder circuit presented here has been designed to enable the highest practicable audio quality to be obtained from a good FM tuner. In particular, the audio signal path has been kept entirely separate from the 38kHz switching circuitry and uses discrete components rather than ICs. The results include a total harmonic distortion figure of 0.02% over the entire VHF/FM signal bandwidth (30Hz–15kHz) in both mono and stereo modes, more than 20dB attenuation of the 19kHz pilot tone and over 60dB attenuation of the 38kHz switching waveform.

These figures are reflected in the overall sound quality provided by the decoder when used in conjunction with a good quality FM tuner. The stereo image position and stereo separation are particularly good. Because of this, the decoder should prove of interest both to those who are building an FM receiver from scratch and those who already have a receiver and wish to upgrade it.

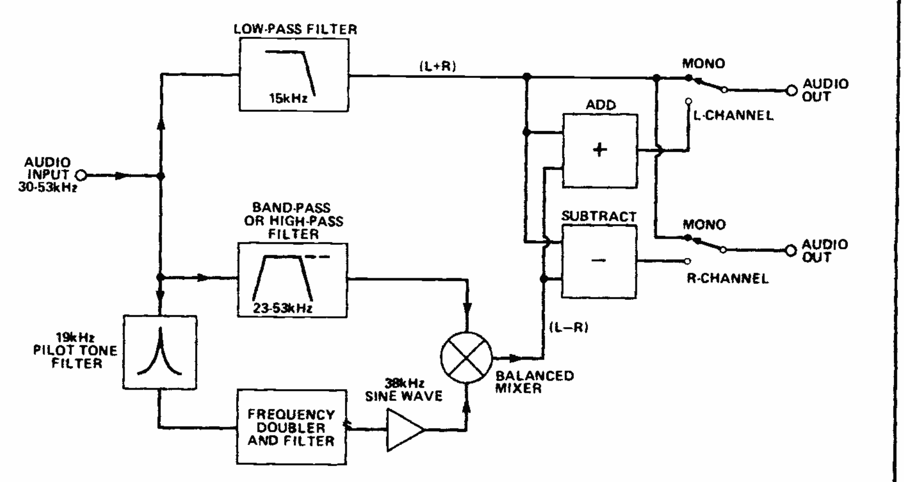


Fig. 1 A matrix-type stereo decoder.

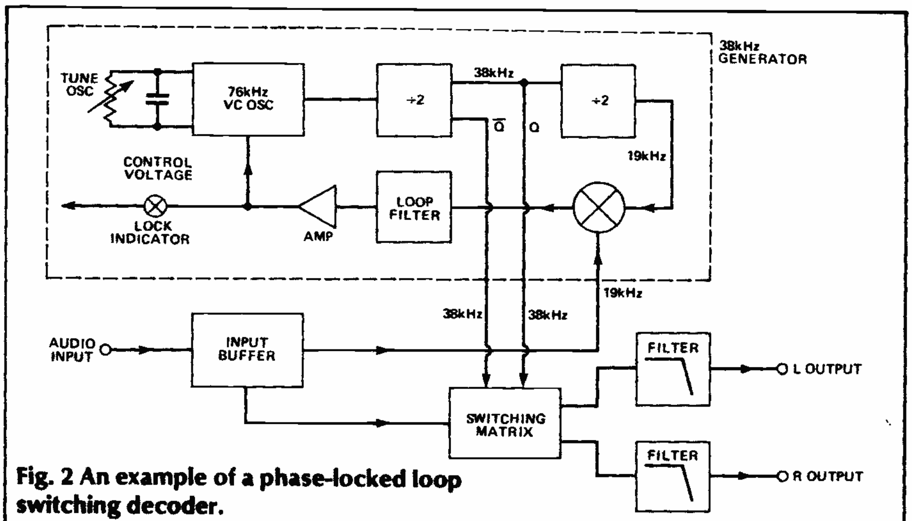


Fig. 2 An example of a phase-locked loop switching decoder.

Stereo Decoding

One of the major advantages of VHF FM radio broadcasting is that the available bandwidth is adequate to allow the transmission of a stereo signal. This is done using the Zenith-GE pilot tone system as described earlier in this series of articles.

The advantage of this method is that the signal is broadcast as a composite. The normal 20Hz–15kHz part of the audio modulation is a perfectly conventional mono (left plus right) transmission, while the necessary additional information required to generate two separate channels is carried as a normally-inaudible suppressed carrier, centred on 38kHz and occupying the 23kHz–53kHz part of the AF spectrum.

There are two basic methods of recovering the stereo information from this composite signal (Figs. 1 and 2). The first of these uses the 19kHz pilot tone to regenerate a 38kHz carrier so that the sidebands can be recovered from the 23–53kHz signal in the normal way. The second method uses a switching system operating at 38kHz, to alternately route the whole signal to the left or the right channel outputs.

Both of these methods effectively do the same thing, but there are differences in the

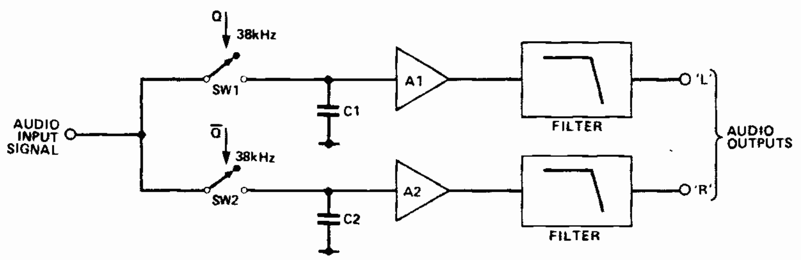


Fig 3. Block diagram of a sample-and-hold signal splitting circuit.

potential signal-to-noise ratio which relate particularly to the effective bandwidth of the decoded signal.

The signal-to-noise ratio of the stereo signal will always be worse than that of the mono signal because it requires more bandwidth. While the magnitude of the signals is the same, the stereo one demands at least a 53kHz bandwidth whereas the mono one only requires 15kHz, and the noise associated with any signal is related to the demodulation bandwidth needed.

However, there is an additional problem with the switching method shown in Fig. 2. If a square wave is used as the switching waveform it will contain a whole series of odd harmonics, at 114kHz, 190kHz, 266kHz etc, and any noise components lying within 19kHz on either side of these will also be commutated down into the audio band.

The matrix system of Fig. 1 is free from this problem — provided that a good quality sine wave is used to replace the 38kHz suppressed carrier. To generate a good sine wave which is accurately locked in phase to the 19kHz pilot tone is not easy, and any error in phase will substantially impair the effective channel separation.

The noise problems associated with the switching decoder can be removed if an effective bandwidth limiting input filter is used to remove any incoming noise signal in the frequency band from 95kHz (38x3–19) upwards. This allows the use of a simpler type of decoder system, which can be built in a form which offers a very low overall distortion of the audio signal.

The Sample And Hold System

Two further requirements of any decoder system, in addition to low noise and low distortion, are that the audio signal output level shall be identical in both mono and stereo modes, and that it is possible to switch from one to the other noiselessly. Both of these additional requirements are met by the sample and hold arrangement shown in Fig. 3.

In the mono mode, when both SW1 and SW2 are closed, it is evident that the same signal is being fed to both the left and right channels. If the switches are sequentially operated, the holding action of capacitors C1 and C2 will result in an output signal which is identical in amplitude to the incoming one, apart from any differences in the left and right signals. This is illustrated for a single channel in Fig. 4.

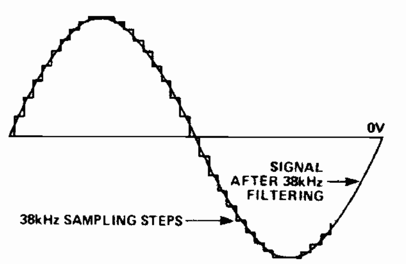


Fig. 4 The sort of waveform produced by a sample-and-hold circuit.

The requirements for correct operation of this circuit are that the switching action should take place at the correct time in relation to the incoming signal waveform, that the time required to charge C1 and C2 from the source through the switching shall be very small in relation to the switching time, and that the discharge time through the input impedances of A1 and A2 shall be very long in relation to the sample period.

In this circuit, Q1 and Q2 form a third-order active low-pass filter. However, only two of the necessary low-pass RC elements are present on the decoder board since the third (5.7us) component was provided by the PLL loop characteristics of the circuit with which it was designed to be used.

Where this decoder is used with a demodulator without AF output low-pass filtering, it is necessary to add the network R41/C27. With the PLL tuner described in the following part of the article, this is switchable to give a wide or narrow effective bandwidth. Where the signal strength is good, leaving SW2 open gives a slightly better image separation. With it closed, the signal-to-noise ratio is improved by about 6dB with only a small loss of stereo width.

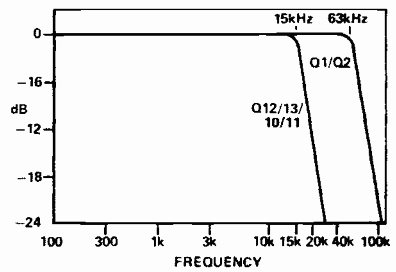


Fig. 5 Transmission characteristics of the input and output filters used in this design.

A similar filter is used after the sample-and-hold switching FETs (Q8, Q9) to produce a rapid attenuation beyond 16kHz. By leaving the 75us de-emphasis network to follow it, the required attenuation of the 38kHz switching waveform ripple is still further increased. The attenuation characteristics of the input and output filter circuits are shown in Fig. 5.

The Switching Waveform Circuit

The 38kHz switching waveform used to drive the FET switches is derived from a standard 1310-type PLL stereo decoder, IC1, which is employed to monitor the incoming signal and extract the 19kHz pilot tone when present.

When the 19kHz pilot tone is present, a 76kHz sawtooth will appear at pin 14 of the phase-locked loop, locked in period to the pilot tone. A DC signal will also be present on pin 6, and can be used to actuate an indicator LED when the 19kHz pilot tone is detected. This DC signal is also used in the circuit shown in Fig. 6 to operate a DC switching transistor, Q4, which controls the supply to IC2. This is a CMOS dual D-type flip-flop, which is used to generate the symmetrical 38kHz switching signal.

In the absence of a DC output current from pin 6 of IC1, Q4 is

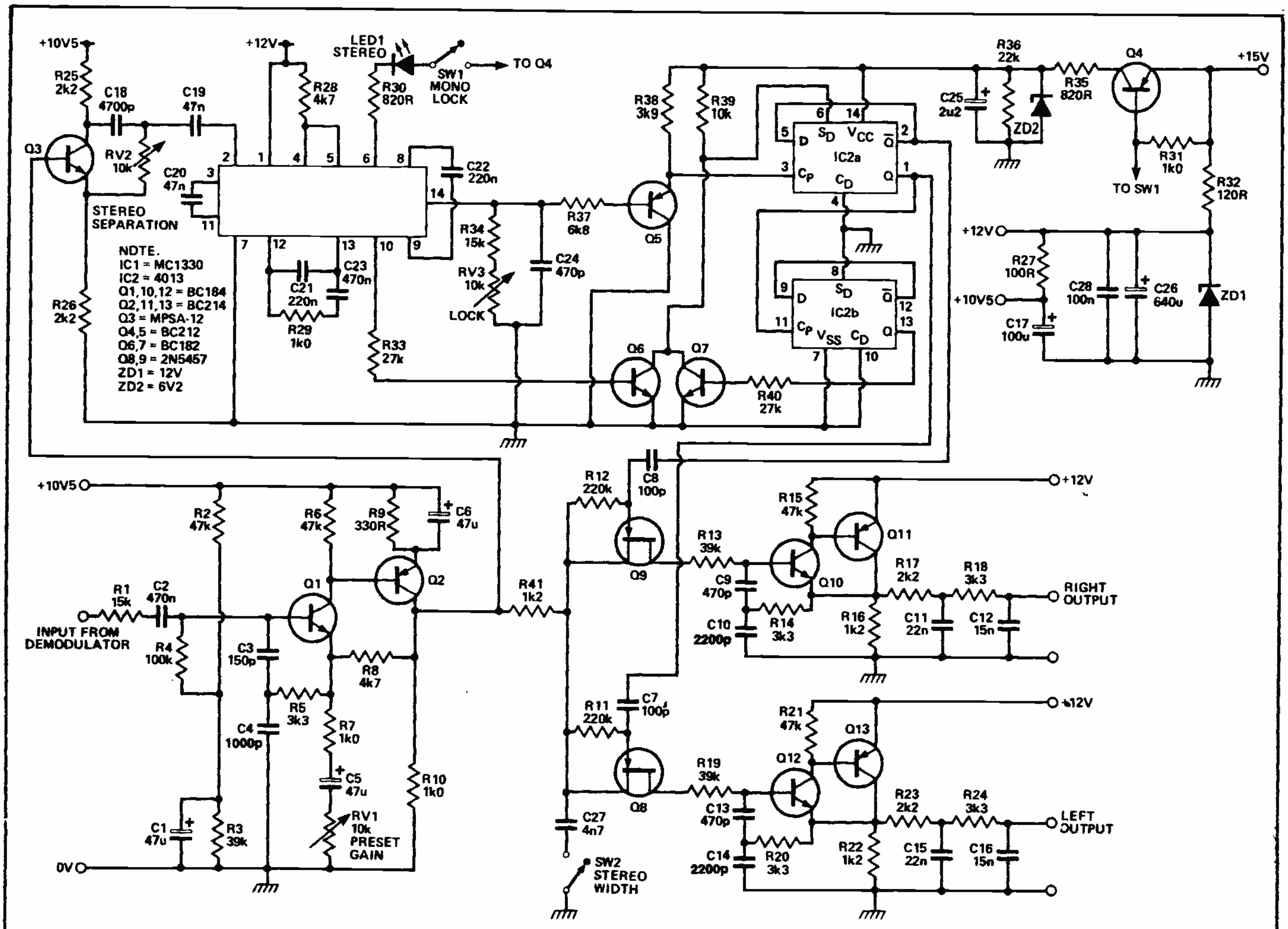


Fig. 6 Complete circuit diagram of the FM stereo decoder.

non-conducting, IC2 is inoperative, and both of the switching FETs (Q8 and Q9) are permanently conducting. The incoming signal from the demodulator will thus be passed directly to both the left and right output channels without interference from the switching/decoding circuitry. Opening switch SW1 will maintain this condition even when a stereo signal is present.

BUYLINES

A complete kit of parts for the FM stereo receiver will be available from Hart Electronic Kits, Penylan Mill, Oswestry, Shropshire SY10 9AF. They will be supplying parts for both the decoder and the other sections of the receiver which have yet to be published, and they will happily provide individual components or part kits as well as full kits. Prices had yet to be finalised at the time of going to press, so intending constructors should contact Hart on (0691) 652 894. The PCBs will be included in the kits, and since Hart will supply these separately too we will not be offering them through our PCB Service.

Provided that a stereo signal is present and the PLL IC has been correctly tuned, a 76kHz sawtooth waveform will be fed to Q5 and to IC2 as a clock signal. The first divide-by-two output signal pair in IC2 produces a 38kHz signal from this clock and drives Q8 and Q9.

Unfortunately, it would be possible if the circuit were switched off for the left and right channels to be reversed, since their identity depends on the relative phase of the 38kHz waveform. This possibility is removed by using the second-stage (19kHz) output from IC2 to feed a simple two-transistor NOR gate, Q6/Q7. The other input to the gate is taken from the 19kHz output of the 1310. If these should become synchronous in phase, the collectors of Q6, Q7 will rise toward the positive rail and reset IC2 for one half cycle, restoring the correct switching phase.

Because of the sample and hold action of the decoder, the stereo/mono transition is completely noiseless.

The best separation between the two stereo channels is given when the phase of the 38kHz switching waveform is timed precisely to coincide with the L-R channel transisitons at the transmitting station encoder, from which the 19kHz pilot tone is derived.

Since, in this circuit, the audio channel and the switching waveform channels are handled separately, it is possible to feed the 19kHz pilot tone through a phase-angle adjusting all-pass filter, Q3. This allows the phase of the pilot tone fed to the PLL (IC1) to be adjusted to give the greatest L–R separation. The adjustment is provided by RV2.

Construction

The majority of the components for the stereo decoder are mounted on a single-sided PCB. The only exceptions are the panel-mounting components (LED1, SW1, SW2, etc) and the second stage of filtering on the two outputs (R18/C12 and R24/C16).

Begin assembling the board by installing the four wire links (all around Q4 and IC2) and the IC sockets if you intend using them. Solder into place the resistors, the presets and the capacitors, then move on to the diodes, the transistors and the ICs. Take the usual care with the electrolytic capacitors and the various semiconductors, all of which must be inserted the right way around if everything is to work correctly.

If you wish to, you can test the audio stages of the board quite simply before fitting it into a receiver. Connect up a +15V supply and apply a signal to the input. You should find that the signal appears at both audio outputs with equal amplitude.

Assuming all is well, mount the board inside the receiver you plan to use it with and make the necessary connections. The off-board filter components, R18/C12 and R24/C16, should be soldered to the far ends of the audio output leads. It will probably be most convenient to mount them on the stereo output sockets.

Switch on the receiver and tune to a station which you know will be broadcasting in stereo. Adjust RV1 to give an audio output level suitable for your amplifier. With SW1 closed, adjust RV3 until the LED lights up and continue turning it until the LED goes off again. Leave RV3 set to the centre of the range over which the LED is on. The outputs of the receiver should now be providing a stereo signal.

Connect a millivoltmeter between the signal connections of the two output sockets and listen for a stereo signal which is fairly random in nature. Something like applause or other crowd noise would be suitable. Set the stereo channel separation by adjusting RV2 until you get a maximum reading on the meter. You should find two maximum positions, one which brings the left and right channels out one way round and one which reverses them. It does not matter which position you leave RV2 at provided you bear in mind that your channel identification may not correspond with that of the broadcasting station.

If you don't have a millivoltmeter, you can set the channel separation roughly by using headphones and adjusting RV2 until the stereo image sounds best.

● *The remainder of the circuitry necessary to produce a complete FM stereo receiver will be described next month.*

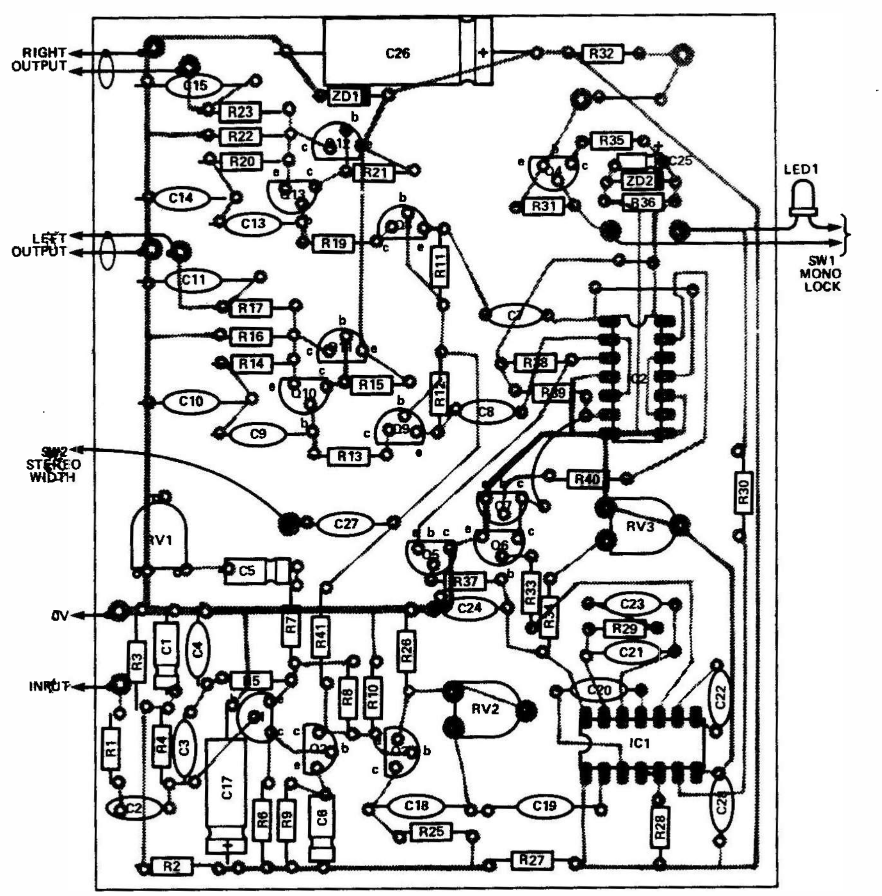


Fig. 7 Component overlay for the stereo decoder PCB. Note that R18/C12 and R24/C16 are not mounted on the board.

PARTS LIST

RESISTORS	
R1	15k
R2, 6, 15, 21	47k
R3, 13, 19	39k
R4	100k
R5, 14, 20, 18, 14	3k3
R7, 10, 29, 31	1k0
R8, 28	4k7
R9	330R
R11, 12	220k
R16, 22, 41	1k2
R17, 23, 25, 26	2k2
R27	100R
R30, 35	820R
R32	120R
R33, 40	27k
R34	15k
R36	22k
R37	6k8
R38	3k9
R39	10k
RV1	10k horizontal skeleton preset
CAPACITORS	
C1, 5, 6	47u 16V axial electrolytic
C2, 3	270n
C3	150p
C4	1n0
C7, 8	100p
C9, 13, 24	470p
C10, 14	2n2
C11, 15	22n

C12, 16	15n
C17	100µ 16V axial electrolytic
C18, 27	4n7
C19, 20	47n
C21, 22	220n
C25	2µ2 16V axial electrolytic
C26	640u 16V axial electrolytic
SEMICONDUCTORS	
IC1	MC1330
IC2	4013
Q1, 10, 12	BC184
Q2, 11, 13	BC214
Q3	MPSA-12
Q4, 5	BC212
Q6, 7	BC182
Q8, 9	2N5457
ZD1	12V 400mW
ZD2	6V2 400mW
LED1	panel-mounting LED
MISCELLANEOUS	
SW1, 2	SPST switches, toggle, push-button, etc as appropriate

PCB; screned cable; IC sockets if required; case, mounting hardware, sockets, 15V power supply, etc, according to application.

ETI

THE ETI LEDLINE ANALYSER

This expandable analyser is the logical choice for testing digital circuits.

The ETI LEDline analyser is a powerful aid to testing complex logic circuits. It is the digital equivalent of a multichannel oscilloscope with storage facilities — at a tiny fraction of the cost!

The circuit diagram for the project is shown in Fig. 1. Each display section consists of a pair of shift registers which clock in logic levels from the device being tested. The levels are displayed on a line of green and red LEDs, representing logic 1 and 0 respectively. The control board contains the trigger circuit, a counter to determine when the displays are full, and a simple interlock to prevent accidental re-triggering.

In use, the clock input of the analyser is connected to the master clock of the board to be analysed. The inputs to the display section are taken from the outputs to be recorded and the trigger is connected to a suitable point to determine the start of the sampling. The switches SW1 and SW2 allow the circuit to be triggered and clocked by either a rising or falling edge, according to the setting.

As an example, suppose you are using a Z80 CPU in interrupt mode 2. In this mode the interrupting peripheral must supply the lower eight bits of an address which points to its service routine. You suspect that the Z80 is receiving the wrong address and would like to check. One way to do this would be to connect the clock of a ten channel version of the analyser to the Z80's clock input, the trigger to $\overline{INT}$ (with SW1 set so that it fires on the falling edge), eight input channels to D_0 – D_7, a further channel to $\overline{IORQ}$, and one to $\overline{M1}$.

When you are ready, press the READY button (what else?) on the analyser and then arrange for the peripheral to send an interrupt. As soon as the interrupt request arrives, the analyser will record the logic levels on the data bus and the two control lines over the next sixteen cycles of the clock. It will then stop and the display will remain steady for you to analyse at your leisure. More about this later. First you've got to build the LEDline analyser!

Construction

Initially you must choose the number of data channels you need. I would suggest four as a minimum. You will need at least ten if you intend to use the analyser for serious work with 8-bit microprocessors. This will allow eight data or address lines and two control signals to be inspected simultaneously.

The display PCB, as supplied by our PCB Service, will record two channels of data. To make a four channel recorder you will need two display PCBs side by side, linked together as shown in Fig. 1. For more channels, just link together as many display boards as you need. If you have a process camera at your disposal (or can talk your local litho printer into doing it for you) you can photograph the foil pattern several times, join the prints together, and then photograph the foil pattern onto transparency to make a master for a single large PCB for the full number of channels you require.

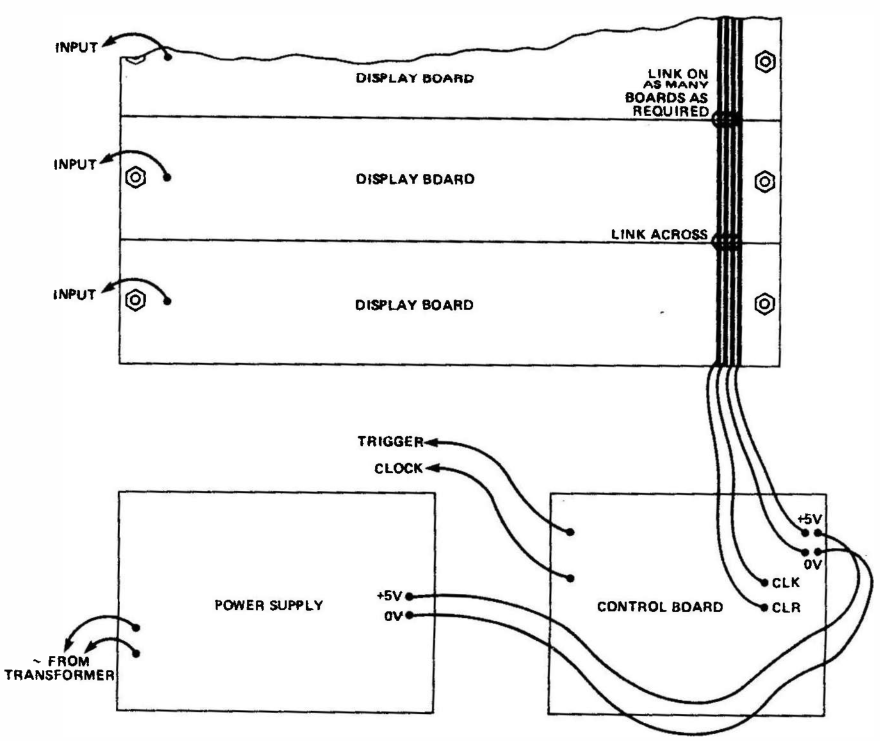


Fig. 1 Internal layout of the case and the board interconnections.

HOW IT WORKS

The display section consists of a pair of eight-bit shift registers, each output lighting a red LED if it is at logic 1 or a green LED for logic 0. As the input is inverted by Q1, the green LEDs indicate a logic 1 input and the red ones a logic 0. Q1 allows the circuit to be driven either by CMOS or TTL. For simplicity, no attempt has been made to match the logic levels exactly. Anything above about 1.5V will be recognised as a logic 1, and anything below will be seen as logic 0. The transistors Q2 and Q3 present a relatively small load to the driving ICs, allowing many more boards to be chained together than would be possible if the clock and clear lines went directly to the inputs of IC1 and IC2.

On the control board, the trigger and clock inputs are once again made compatible with either CMOS or TTL by Q4 and Q7 respectively. IC3a allows the analyser to be clocked by either the rising or falling edge of the input (selected by SW1). IC3b and SW2 perform the same function on the clock input. Transistors Q5 and Q6 and associated components perform a power-on reset function: the circuit powers up with all the displays at 0, IC5 set to zero, and IC4a and IC4b set to 1. In this condition, the circuit will ignore a trigger input, since IC4a is already preset.

Pressing the ready button clears IC4a, and clears down the rest of the circuit as for a power-on reset. It also lights LED33 to show that the analyser is ready for action. Once IC4a is cleared, a rising edge at its clock input (from the trigger) will cause it to change state. When this happens it will be returned to the preset state and will ignore any further trigger inputs until the ready switch is pressed again. LED33 will then be extinguished to show that the circuit has been triggered. IC4b will receive a rising edge at its clock input via IC3c and will change state, enabling the clock input via IC3d. This sets the sampling process in motion. All the display boards are clocked via IC3d and collect data from the board under test.

On the fifteenth clock pulse, pin 15 of IC5 will go high. On the sixteenth it will go low again, at this point clocking IC4b via IC3c. This causes the clock pulses to be cut off and the circuit remains in this state, with sixteen sets of samples displayed, until SW3 is pressed to start another cycle.

BUYLINES

The low current LEDs used for this project were obtained from Electromail, stock no. 588-386 and 588-392. The test probes were from the same source, stock no. 424-175, as was the heatsink for the regulator (stock code 402-989. Note this item comes in packs of 10). In case of difficulty, the 15,000μ capacitor is also available from Electromail (104-348). The PCBs will be available from our PCB service in due course. None of the other components should present any difficulty.

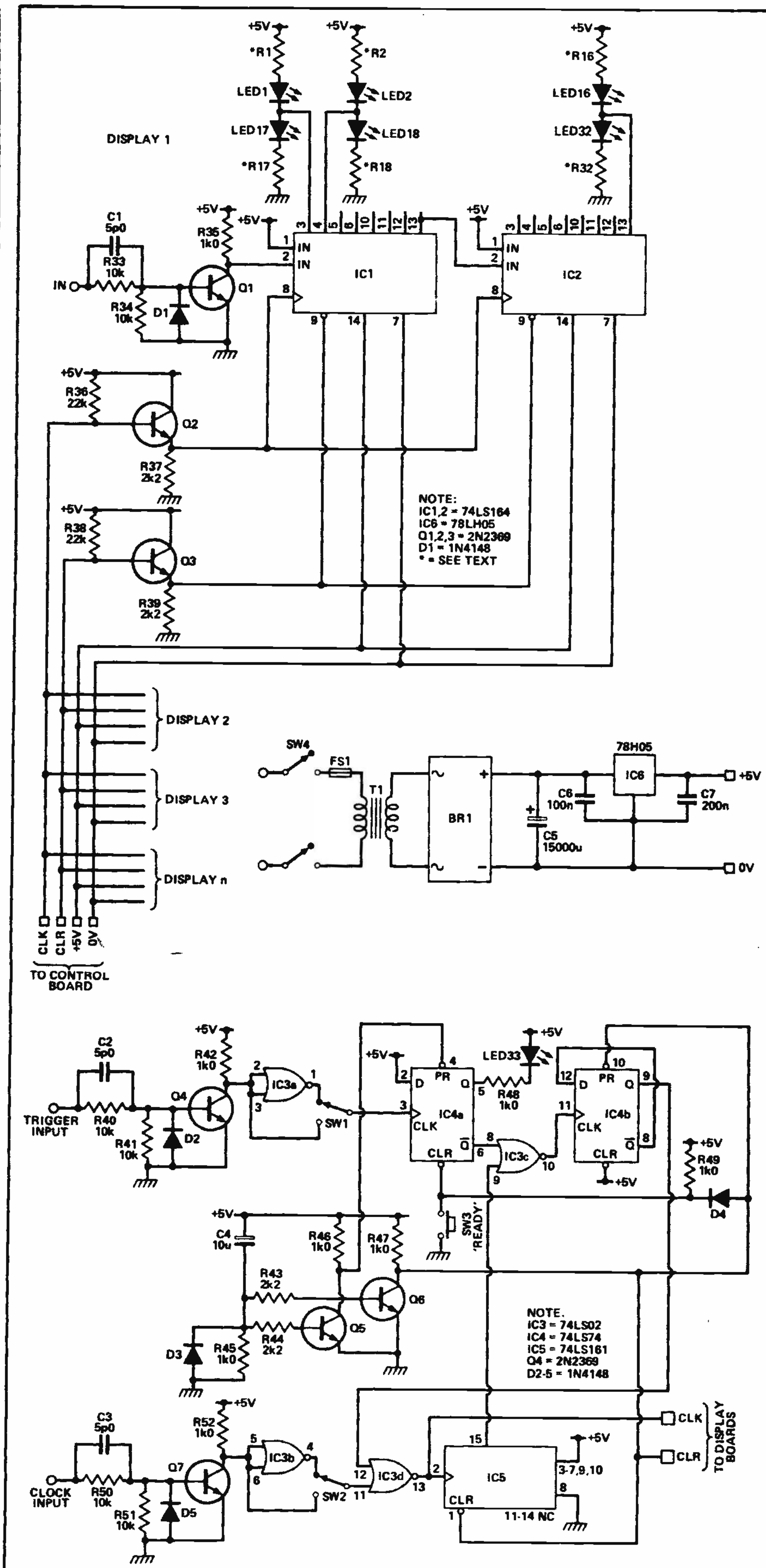
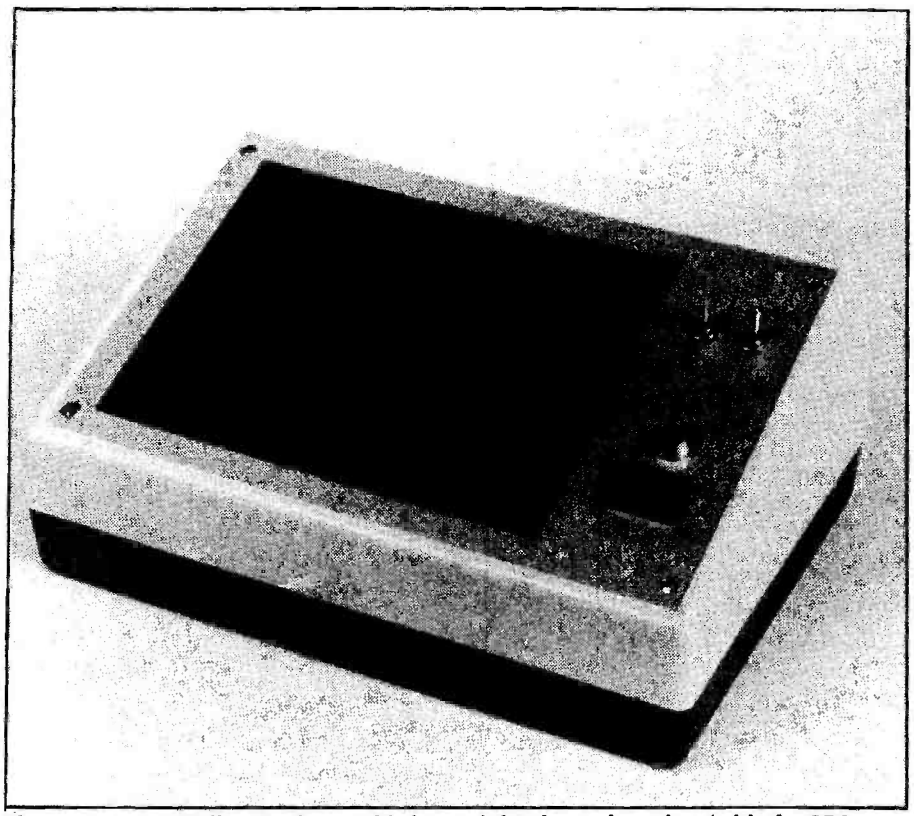


Fig. 2 The circuit of the LEDline Analyser.

The component overlays for the PCBs are shown in Fig. 3. For the display LEDs we used high efficiency, low current types which give a good output at 2mA. If you are building an anlayser with only four or so channels, you may prefer to use lower cost LEDs (in which case R1 to R32 must be reduced to 330 ohms). However, for six or more channels, it would be advisable to use the high efficiency types to avoid straining the power supply!

To economise even further, the LEDs 17–32 and the associated resistors R17–32 can be omitted. This would leave just one row of LEDs for each channel. Low logic levels would then be indicated by an unlit LED in the remaining row.

For the analyser inputs you can either have a separate socket on each board (the PCB will accommodate a miniature BNC socket or a 2mm 'banana' socket) or else bring all the inputs to a single connector and make the test connection via a length of ribbon cable. In either case, keep the test leads as short as possible, and pay particular attention to the ground connection.



The prototype LEDline Analyser. This is an eight channel version (with the 256 LEDs behind the central filter on the front panel). The connections to the unit under test are from the rear panel.

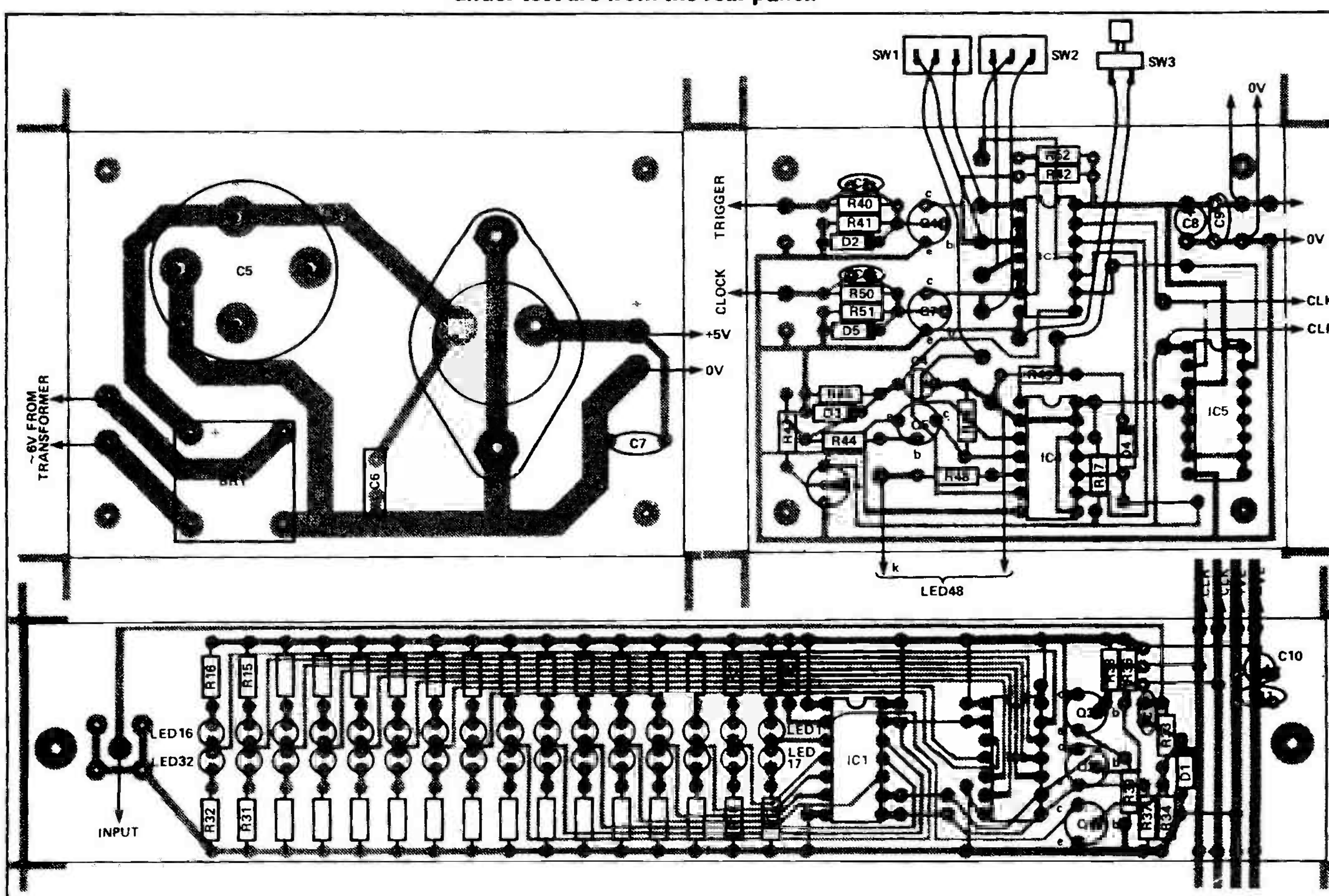


Fig. 3 Component overlays for the LEDline analyser. (a) the power supply board, (b) the control board, (c) a single display channel.

Using The Analyser

A good earth connection between the analyser and the circuit to be tested (the 'unit under test', or UUT, as the ATE trade would have it) is essential, particularly for logic with a clock speed of 1MHz or above. This means keeping the connection short, using relatively heavy gauge wire (32/0.2) and making firm connection to a suitable grounding point on the UUT. Connection test leads to IC pins can be made with a sprung connector such as type 424–175 from Electromail.

I have already mentioned how the analyser might be used to investigate a Z80 interrupt. With this IC, an interrupt acknowledge cycle is identified by both $\overline{\text{M1}}$ and $\overline{\text{IORQ}}$ going low. Having triggered the analyser from $\overline{\text{INT}}$ and captured the data, the next step would be to scan along the line of LEDs until you find the 0 LED lit for both $\overline{\text{M1}}$ and $\overline{\text{IORQ}}$. If the analyser has been clocked from the rising edge of the Z80's clock, you will have found point 5 of Fig. 4 — the rising edge of the clock on the second 'wait' cycle. The very next column of LEDs along (captured on the rising edge of T_3) will give you the bits the Z80 will take to be the lower byte of the interrupt vector. If these are not what you intended, you've found the fault!

However, note that the acknowledge sequence may vary. Extra 'wait' states could be inserted by the hardware, for example.

When using the analyser with asynchronous or 'event driven' logic there is no obvious signal available to connect to the analyser's clock. In fact it is probably better to regard the clock input as the 'sample time' input — even on synchronous logic. If we stick to the Z80 for a moment, a better way to sample the interrupt data might be to use $\overline{\text{IORQ}}$ to clock the analyser. Once again, the trigger could come from $\overline{\text{INT}}$, and data would be clocked in whenever $\overline{\text{IORQ}}$ went high, so up to 16 interrupts could be sampled. Of course, you would have to be sure the Z80 was not performing any other I/O routines in between, otherwise these would be recorded too. The same principle applies to testing asynchronous logic — find a signal which marks the event you wish to record, and use this to clock the analyser.

Improvements

The analyser was designed to be as simple as possible, consistent with performing a useful function. The Readers' Survey has shown many readers like to modify ETI circuits (How dare you!), so here are a few ideas.

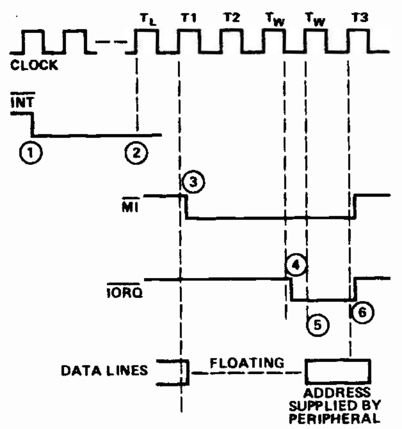
An improved triggering function could be an advantage. To return yet again to Fig. 4, it could be useful to trigger the circuit only when both $\overline{\text{M1}}$ and $\overline{\text{IORQ}}$ were low. With an extra trigger input and logic gate, this could be arranged without difficulty. An OR triggering function is another possibility — it may sometimes be desirable to set the analyser going when either one of two different events occur.

With intermittent faults, it might be useful to detect the occurrence of the fault and to know the logic states which led up to it. In this case, the analyser could sample continuously and the trigger would be used to *stop* the recording rather than to start it. When the fault condition appeared, the analyser would then freeze the preceeding 16 logic states which could give a clue to the cause of the fault.

Yet another possibility would be a delay in the clock line. If you wanted to record the data clocked into a register, it may be an advantage to clock the analyser from the register's clock input but to record when the data has had a chance to settle at the register's outputs.

There are numerous other possibilities — don't forget our Tech Tips section if you come up with anything good!

ETI



Note: At point 1 the peripheral takes the Z80 interrupt line low. The Z80 ignores the interrupt until the final clock cycle of the current instruction (T_L) when, at point 2, it tests the interrupt line and prepares for an interrupt acknowledge cycle. At point 3 $\overline{\text{M1}}$ goes low (as it would for the opcode fetch which would usually follow the completion of an instruction) and 2½ clock cycles later (point 4) $\overline{\text{IORQ}}$ goes low, identifying the cycle as an interrupt acknowledge. At point 6 the bits present on the data lines will be read into the Z80 and used as a pointer to the interrupt service routine. This is the critical point for deciding whether the peripheral is supplying the correct data.

Fig. 4 Simplified timing diagram of the Z80 interrupt acknowledge, mode 2.

PARTS LIST

CONTROL BOARD AND POWER SUPPLY

RESISTORS (all ¼W, 5%)

R40, 41, 50, 51	10k
R42, 45, 46, 47, 48, 49, 52	1k0
R43, 44	2k2

CAPACITORS

C1, 3	5p ceramic
C4	10µ 6V tantalum
C5	15,000µ 16V electrolytic
C6	100n polyester
C7	200n polyester
C8, 10	10µ tantalum
C9	100n Cer.

SEMICONDUCTORS

IC3	74LS02
IC4	74LS74
IC5	75LS161
IC6	78H05
Q4–7	2N2369
D2–5	1N4148
BR1	200V, 6A bridge rectifier

MISCELLANEOUS

Transformer (6V, 20VA), fuse, fuse holder, mains switch, heat sink or regulator, case, PCBs, test leads and connectors, pushbutton and 2 x SPDT switches.

DISPLAY BOARD

One set of the components listed below will be required for each channel.

RESISTORS (all ¼W, 5%)

R1–32	1k0 (see text)
R33, 34	10k
R35	1k0
R36, 38	22k
R37, 39	2k2

CAPACITORS

C1	5p ceramic
C10	10µ 6V tantalum
C11	100n ceramic

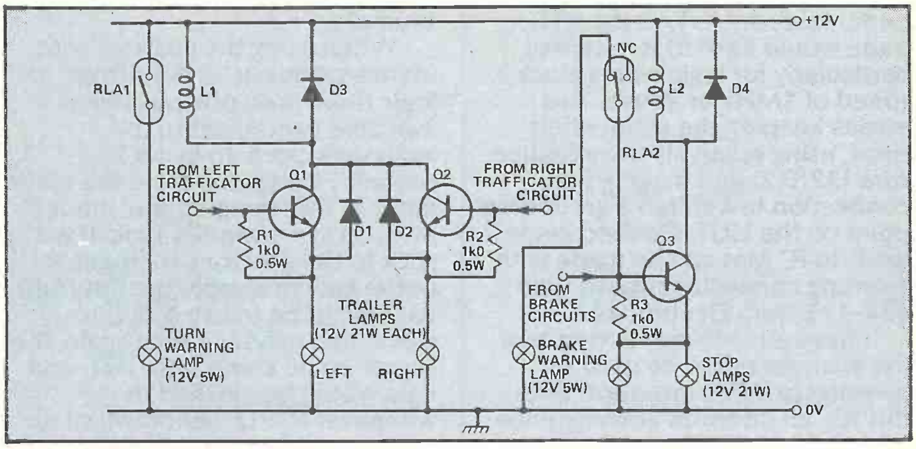
SEMICONDUCTORS

IC1, 2	74LS164
Q1–3	2N2369, 2N2222, or similar high speed switching transistor
LED1–16	High efficiency green LED
LED17–32	High efficiency red LED
D1	1N4148

TECH TIPS

Trailer Light Controller

P. J. Dinning
Newcastle-upon-Tyne



A trailer's lights put a heavy demand on the towing car's system and can change the flash rate of indicators, burn out the flasher unit or damage the brake light switch. This circuit will allow extra indicator and brake lights to be safely added to a car's electrical system, typically drawing only 4% extra current.

The circuit also provides for a pilot light to indicate correct functioning of the flashers and another to warn of bulb failure in the brake light circuit.

TIP35 transistors are specified, but any plastic power type is suitable provided it has a collector current rating of at least 15A. This is because bulbs which consume 1.75A when lit have cold surge currents around five times greater. Darlington transistors should not be used.

The base-emitter resistors (R1 and R2) are included because of the high temperatures experienced in cars and trailers. They prevent the transistor turning on due to leakage.

Freewheel diodes D1 and D2 prevent voltage spikes damaging the transistors.

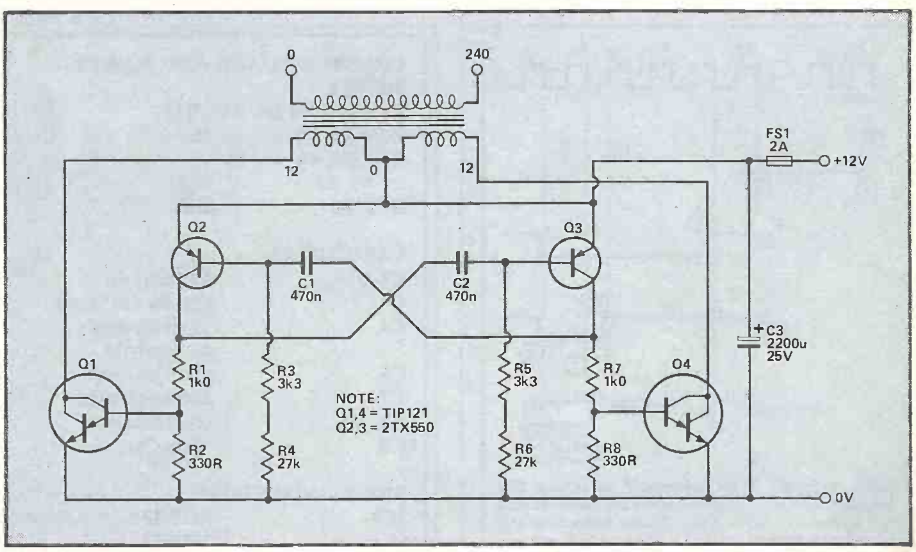
The relays RLA1 and RLA2 are reed switches. In the prototype, an Electromail type 338-147 was used for the flasher circuit and type 338-888 for the brake circuit. These are rated at 40 Ampere turns to operate. With one flasher or brake light bulb the coils L1 and L2 should be 23 turns of 1mm copper wire. For two bulbs 12 turns should be enough. The coil should be wound in one or two layers around the centre of the reed switch and secured with a rubber adhesive (epoxy glues should not be used as the reed switches are fragile).

Stop light and flasher bulbs use a common return system through the car's bodywork and many weird effects can be traced to poor or rusted earth connections. This should be born in mind when fault finding.

The unit can be easily constructed in a small diecast box, which would also provide heatsinking for the transistors. The TIP35 tabs are connected to the collector and should be insulated with bushes and mica washers. If a steel box is used the reeds should be mounted at least 1" away from the sides.

50Hz Inverter

Andy Armstrong
Leighton Buzzard



This circuit uses an ordinary mains transformer in reverse to give an output of about 220VAC at approximately 50Hz. A toroidal transformer, such as the ILP 30VA type, is suitable.

Darlington power transistors are used to drive the secondary of the mains transformer with a 50Hz square wave. This type of transistor includes an internal anti-parallel diode which catches any spikes due to leakage reactance. Any constructors wishing to make their own Darlingtons from discrete transistors should add diodes with the cathode towards the transformer to fill this function.

The base drive for the power transistors is provided by a conventional multivibrator. The component values are shown to give operation at just over 50Hz. It is better to run the circuit at just over rather than just under because at lower frequencies there is a danger that the transformer will saturate.

The choice of Q2 and Q3 is important to ensure that the circuit operates reliably, because it is heavily loaded. In some cases BC212s will work but this cannot be relied upon. The circuit will not work reliably unless the connections between the transformer centre tap, the emitters of the power transistors and C3 are all kept short.

This circuit has proved useful as a shaver adaptor to enable me to shave on the way to work while stuck at the traffic lights. There is always enough traffic holdup to allow time for shaving and it means a few minutes more sleep in the morning. This is probably as dangerous as using a telephone on the move, but the unit might be useful for camping.

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E8206-5	Logic Lock	F
E8208-1	Playmate Practice Amp 3bds SA1	K
E8212-1	ELCB	F
E8301-2	Analogue to digital conv (ZX81/ Spectrum)	E
E8305-3	Dual Audio Power Supply, Linsley Hood	G
E8305-5	Balanced Input Preamplifier	F
E8307-2	Flash Trigger-sound or FR	F
E8308-1	Graphic Equaliser ⅓ Oct/Chnl	M
E8308-2	Servo Fail-safe	C
E8309-1	NICAD Charger/Regenerator	F
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E8311-1	Mini Drum Synth	F
E8311-8	Moving Coil Pre-Preamp	F
E8312-3	Light Chaser EPROM Controlled (2 Boards)	K
E8402-1	Speech Board	M
E8402-2	Modular Pre-amp Disc Input Mono	F
E8402-3	Modular Pre-amp Stereo Output	F
E8402-4	Modular Pre-amp Relay, PSU	F
E8402-5	Modular Pre-amp Tone Main Mono	F
E8402-6	Modular Pre-amp Tone Filter, Stereo	F
E8402-7	Modular Pre-amp Balanced Output	F
E8402-8	Modular Pre-amp Headphone Amp	F
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E8405-1	Auto Light Switch	F
E8405-2	ZX81 EPROM Programmer	N
E8405-3	Mains Remote Control Transmitter	H
E8405-4	Centronics Interface	F
E8405-6	Drum Synth	F
E8406-1	Oric EPROM Board	O
E8406-2	Spectrum Joystick	E
E8406-3	Audio Design RIAA Stage	G
E8406-4	AD Buffer/Filter/Tone	H
E8406-5	AD Headphone Amp	F
E8406-6	AD Preamp PSU	K
E8406-7	AD Power Amp	H
E8406-8	AD Power Amp PSU	J
E8406-9	AD Stereo Power Meter	F
E8406-10	AD Input Clamp	C
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E8408-2	EPROM Emulator	N
E8408-3	Infrared Alarm Transmitter	E
E8408-4	Infrared Alarm Receiver	F
E8409-1	EX42 Keyboard Interface	F
E8409-2	Banshee Siren Unit	F
E8410-1	Echo Unit	F
E8410-2	Digital Cassette Deck	N
E8410-3	Disco Party Strobe	H
E8411-5	Video Vandal (3 boards)	N
E8411-6	Temperature Controller	D
E8411-7	Mains Failure Alarm	D
E8411-8	Knite Light	D
E8411-9	Stage Lighting Interface	F
E8411-10	Perpetual Pendulum	E
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E8412-6	Active - 8 LF EQ	F
E8412-7	Active - 8 Equaliser	F
E8501-3	Digital Delay (2 boards)	T
E8502-1	Digital Delay Expander	N
E8502-2	Data Logger	J
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E8503-3	THD Meter Mains PSU	F
E8504-1	Framestore Memory	M
E8504-3	Framestore Control	N
E8504-4	Buzby Meter	E
E8504-5	CCD Delay	F
E8505-5	Stereo Simulator	F
E8506-1	Audio Mixer Main	J
E8506-2	Audio Mixer PSU	F
E8506-3	Audio Mixer RIAA	D
E8506-4	Audio Mixer Tone Control	D
E8506-5	EPROM Prog MKII	O
E8508-1	RCL Bridge	N
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E8508-3	EPROM Emulator	L
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E8511-5	Chorus Effect	H
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E8511-8	Switching Regulator	E
E8511-9	Second Line of Defence	M
E8512-1	Specdrum connector	F
E8512-2	MTE Pulse Generator	H
E8511-3	Specdrum	L
E8601-2	Walkmate	L
E8601-3	MTE Counter-timer	M
E8602-1	Digibaro	O
E8603-2	Programmable Logic Evaluation Board	H
E8603-3	Sound Sampler Analogue Board	R
E8604-1	JLLH PA PSU	H
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E8606-2	Midi-CV Converter PSU	D
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BOARDS AVAILABLE

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E8610-3	Audio Analyser Display	H
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E8611-1	Audio Switcher (2 bds)	H
E8611-2	PLL Frequency meter (4 bds)	Q
E8611-3	Upgradeable Amp PSU	J
E8611-4	Call meter, main bd.	O
E8611-5	Call meter, interface bd	N
E8612-1	Bongo Box	J
E8612-2	Biofeedback monitor (Free PCB)	E
E8701-1	RGB Converter	F
E8701-2	Mains Controller	D
E8701-3	Flanger	H
E8701-4	Audio Selector main board	M
E8701-5	Audio Selector PSU	H
E8701-6	Tacho-Dwell	F
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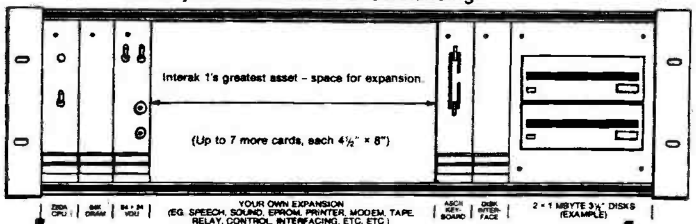

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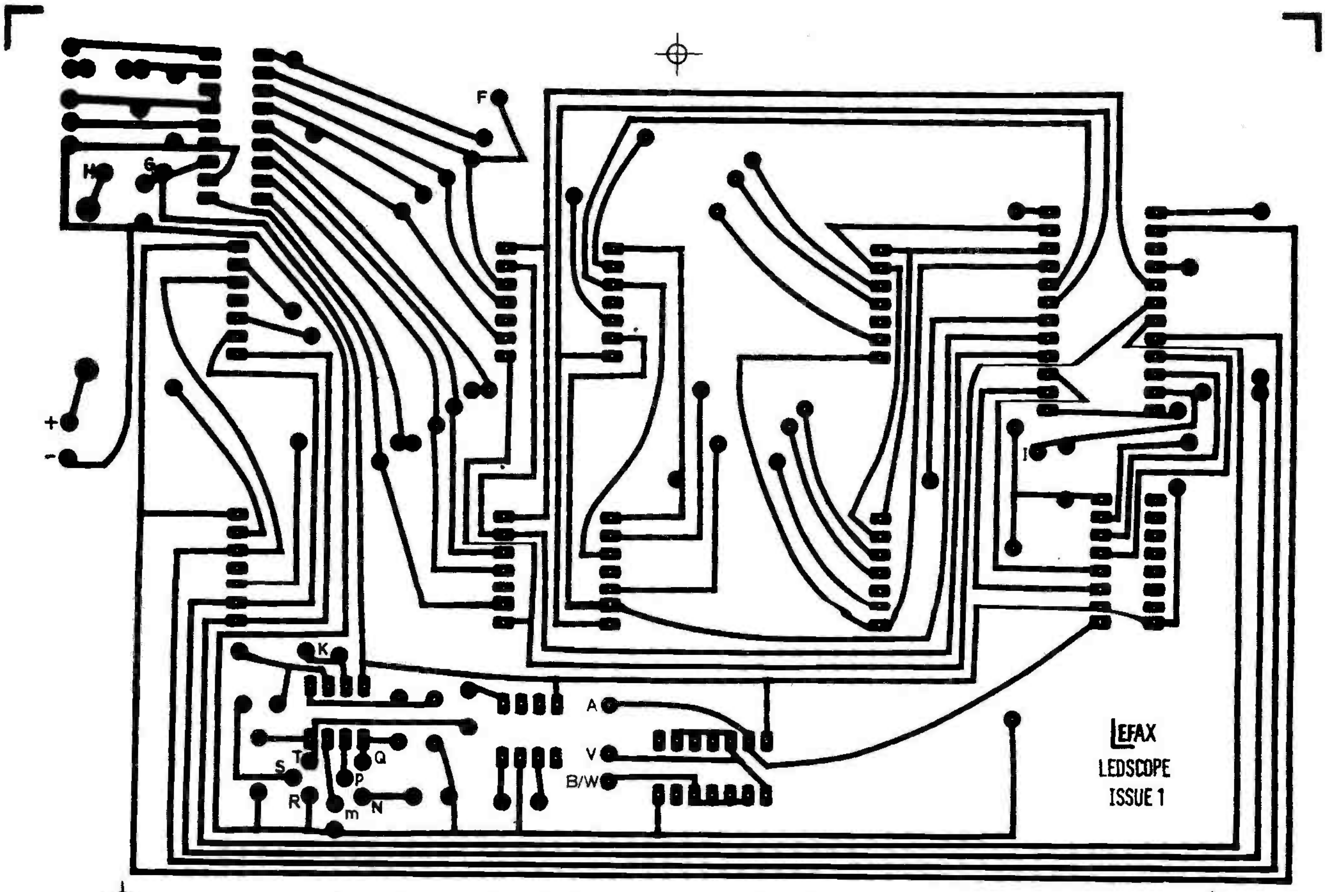
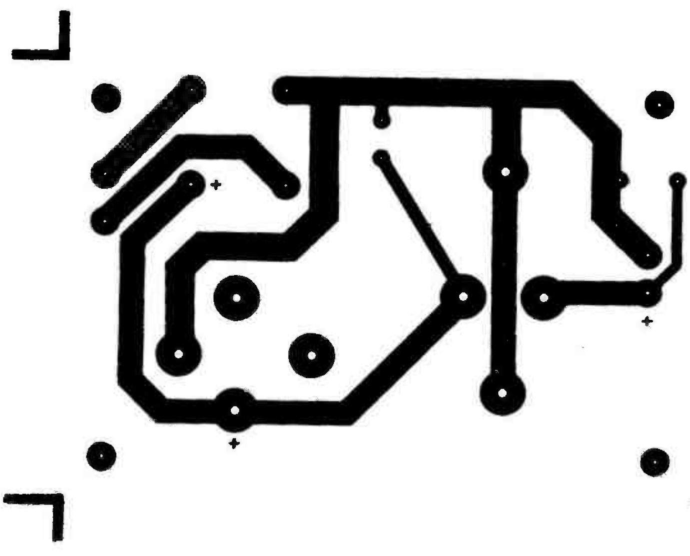
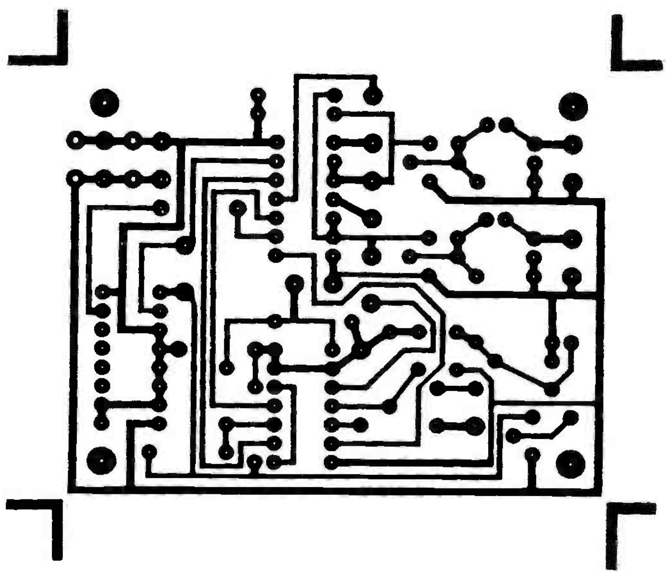


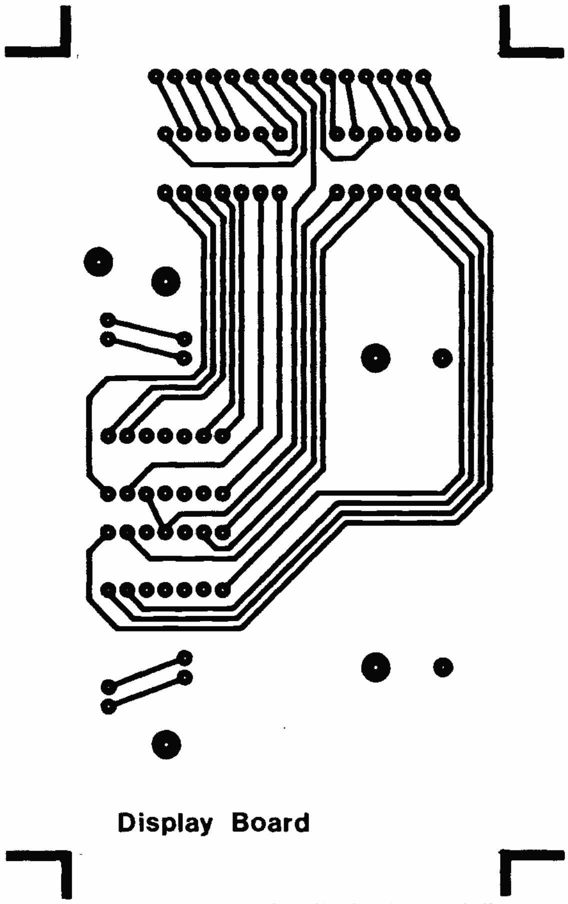
The foil pattern for the LEDscope board, held over from last month.



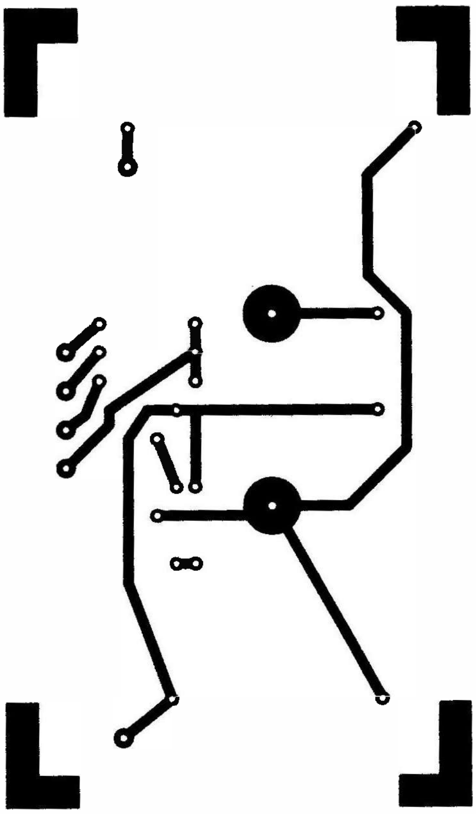
The LEDline PSU board.



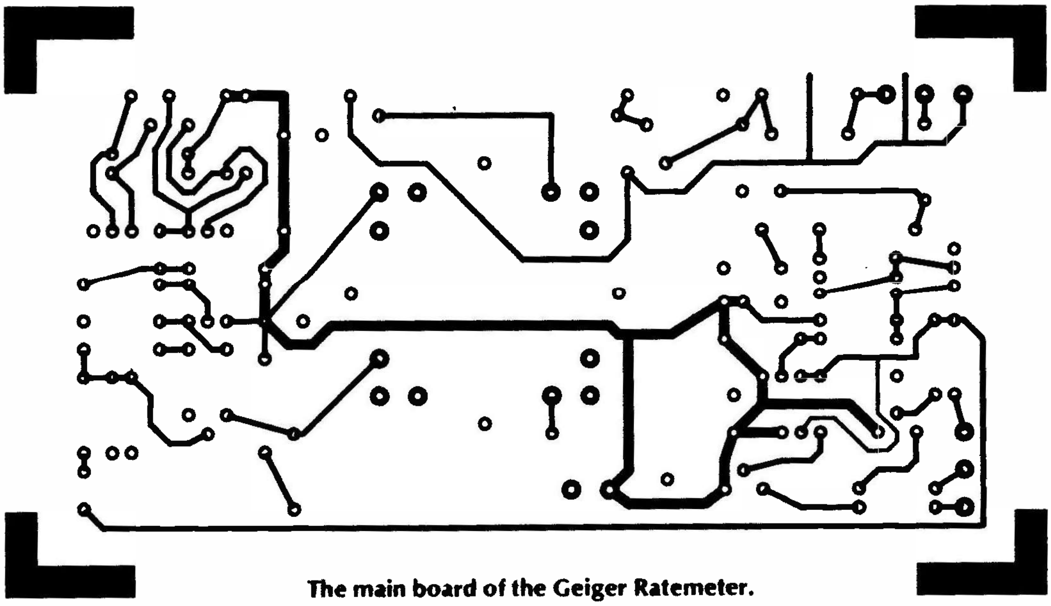
The controller board for the LEDline.



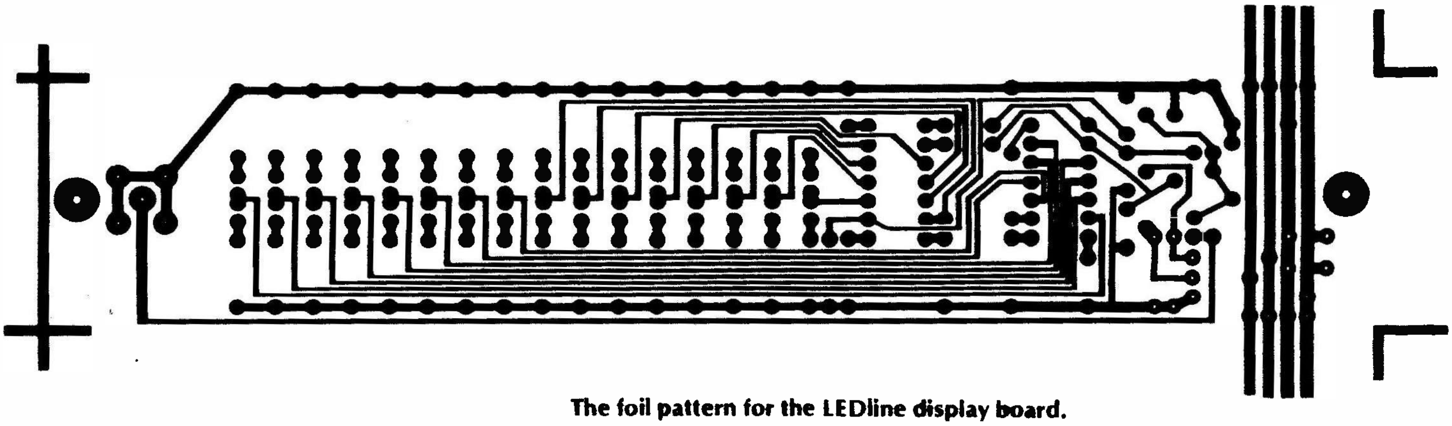
The Photo Process Controller display board foil pattern.



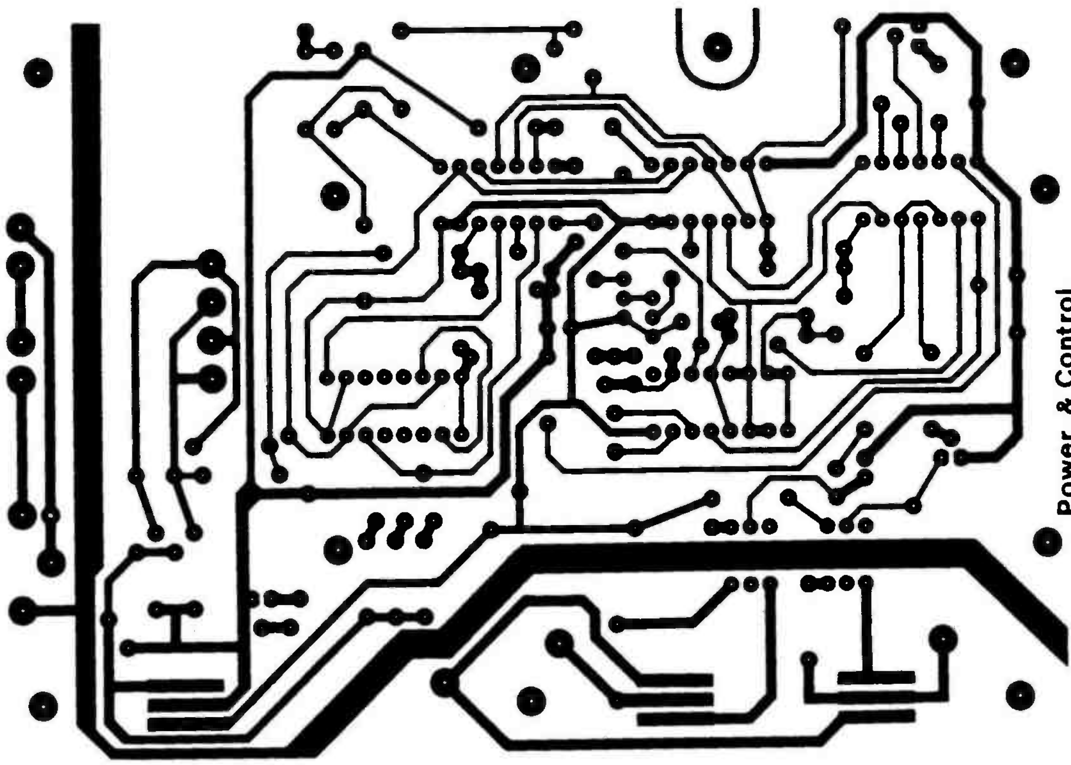
The Geiger Ratemeter ranging board.



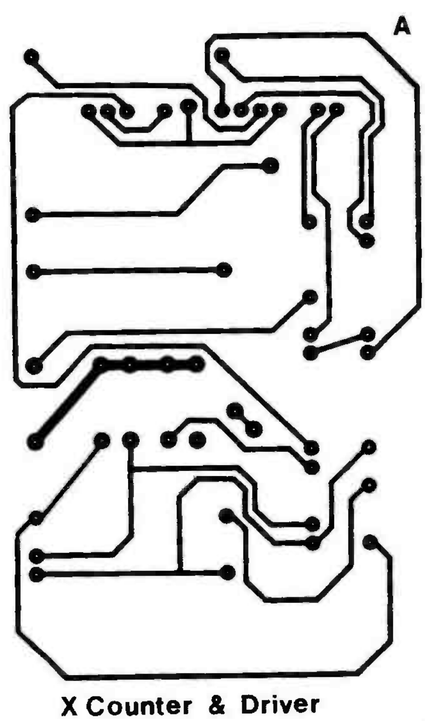
The main board of the Geiger Ratemeter.

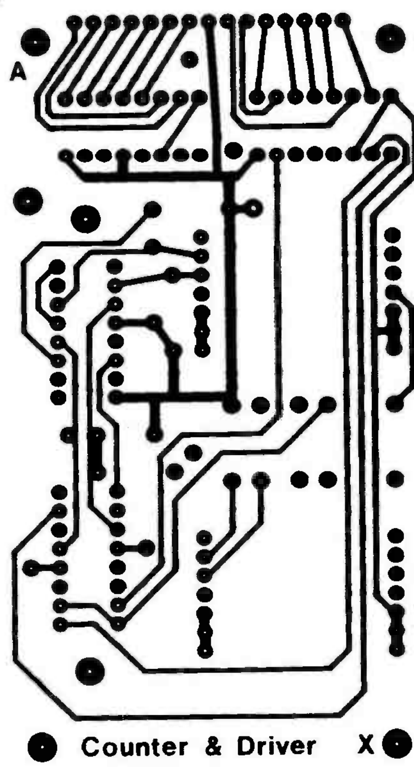


The foil pattern for the LEDline display board.



The main board of the Photo Process Controller.





The top and bottom foils for the Photo Process Controller counter board.

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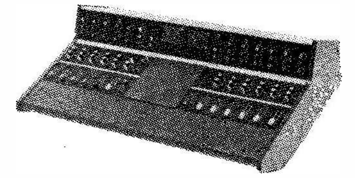

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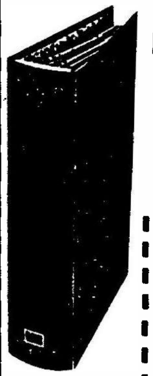

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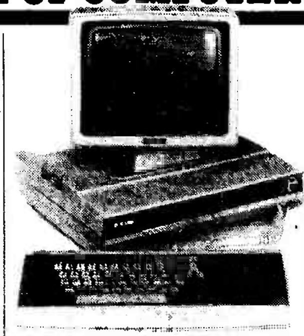

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